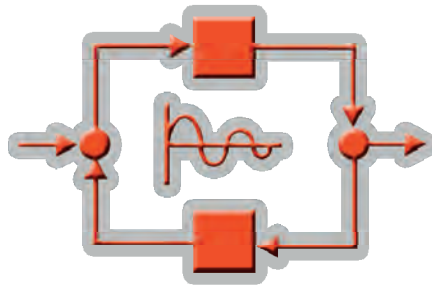




The International Congress for global Science and Technology



ICGST International Journal on Automatic Control & System Engineering (ACSE)

**Volume (11), Issue (I)
June 2011**

**www.icgst.com
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www.icgst-ees.com**

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ACSE Journal
ISSN: Print 1687-4811
ISSN Online 1687-482X
ISSN CD-ROM 1687-4838
© ICGST LLC, Delaware, USA, 2011

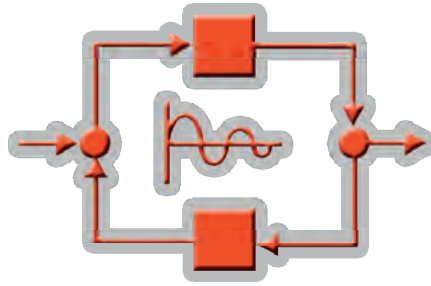
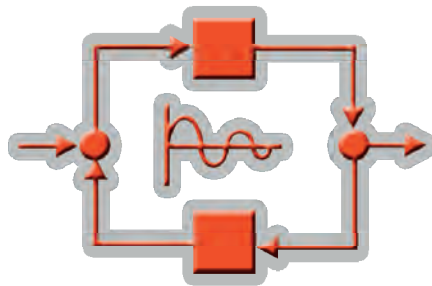


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**ICGST International Journal on Automatic Control & System
Engineering - (ACSE)**

**A Publication of the International Congress for global
Science and Technology - (ICGST)**

ICGST Editor in Chief: Ashraf Aboshosha

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editor@icgst.com

BibTex

@ARTICLE {P1111044323,
AUTHOR = {S.Ravi and P.A.Balakrishnan},
TITLE = {GENETIC ALGORITHM BASED TEMPERATURE CONTROLLER FOR
PLASTIC EXTRUSION SYSTEM},
JOURNAL = {ICGST International Journal on Automatic Control and Systems
Engineering, ACSE},
YEAR = {2011},
MONTH= {June},
VOLUME = {11},
ISSUE = {1},
PAGES= {1--8},
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JOURNAL = {ICGST International Journal on Automatic Control and Systems
Engineering, ACSE},
YEAR = {2011},
MONTH= {June},
VOLUME = {11},
ISSUE = {1},
PAGES= {9--14},
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converters based upon an indirect conversion scheme which models the Matrix
Converter (MC) as two independent stages performing rectification and inversion.
The performance evaluation of 3 Φ to 3 Φ matrix converter using indirect space
vector modulation (ISVM) algorithm is implemented for the Induction Motor
drive. The simulation results are presented which validates the modulation
technique and operation of Matrix Converter (MC)},
NOTE={Matrix converter, AC-AC Converter, Indirect Space Vector modulation,
Performance Evaluation, Induction Motor}}

@ARTICLE {P1111112536,
 AUTHOR = {Ayman A. Aly and S. Elnaggar},
 TITLE = {Genetic PD Control of a Three-Link Rigid Spatial Manipulator},
 JOURNAL = {ICGST International Journal on Automatic Control and Systems Engineering, ACSE},
 YEAR = {2011},
 MONTH= {June},
 VOLUME = {11},
 ISSUE = {1},
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 NOTE= {PD Control, Modeling, Rigid Manipulator, Genetic algorithms} }

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 AUTHOR = {Amara SAIDI, Hsan HADJ ABDALLAH},
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 JOURNAL = {ICGST International Journal on Automatic Control and Systems Engineering, ACSE},
 YEAR = {2011},
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 ABSTRACT= {The rate of dissipation of transient energy is usually used as a tool to measure dynamic system damping. For such a given system equipped with flexible alternative current transmission devices, the above concept is then applied to determine the additional damping. This paper applies and discusses this control strategy for suppressing undesirable electromechanical oscillations in power system with both static synchronous series compensator and static synchronous compensators. The same devices are studied by applying a new method of injection model. Dynamic model of these devices are derived for the two control strategy. Several disturbances are studied for the single and multimachine power systems. It is found that the found results convey eminent efficiency of the described theories. MATLAB/SIMULINK software is used for all simulations},
 NOTE= {SSSC , STATCOM, transient energy function, damping} }

@ARTICLE {P1111117557,
 AUTHOR = {K.Subbaramaiah and V.C.Veera Reddy},
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 MONTH= {June},
 VOLUME = {11},
 ISSUE = {1},
 PAGES= {35--42},
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 NOTE= {Fuzzy Logic Controller, Static Synchronous Series Compensator, TCPS, Deregulation}}

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 MONTH= {June},
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 ISSUE = {1},
 PAGES= {43--50},
 ABSTRACT= {Through the deterministic SVPWM algorithm gives good performance, it generates more acoustical noise. Hence, to mitigate the acoustical noise of the drive, this paper presents an efficient variable delay random pulse width modulation (VDRPWM) algorithm for direct torque controlled induction motor drive. The proposed VDRPWM technique is developed by using the notion of imaginary switching times, which are proportional to the instantaneous phase voltages. Hence, the proposed algorithm did not use angle and sector information and hence reduces the computational effort. Hence, the implementation of the proposed algorithm is simpler. Moreover, the proposed algorithm randomizes the switching periods by varying the delay of switching cycles with respect to corresponding sampling cycles. This randomization of switching periods has a mitigating effect of the acoustic and electromagnetic noise emitted by the supplied system. To validate the proposed PWM algorithm, simulation studies have been carried out and results are presented. The simulation results confirmed the features of proposed PWM algorithm},
 NOTE={acoustic noise, direct torque control, imaginary switching times, space vector pulsewidth modulation (SVPWM), variable delay random pulsewidth modulation (VDRPWM)}}



GENETIC ALGORITHM BASED TEMPERATURE CONTROLLER FOR PLASTIC EXTRUSION SYSTEM

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Abstract

Temperature control of plastic extrusion system suffers from problems related to undesirable overshoot, longer settling time and disturbances. PID tuning plays an important role in the complex process such as the temperature control of a plastic extrusion system. A number of temperature control process can be done using PID controllers. Most of the process is complex and nonlinear in nature resulting into their poor performance when controlled by traditional tuned PID controllers. Determination and tuning of the PID parameters continues to be important, as these parameters have a great influence on the stability and performance of the control system. This paper proposes Genetic Algorithm to improve the performance of the temperature control in plastic extrusion system. The methodology and efficiency of the proposed method are compared with that of traditional methods and the results obtained reflect that use of Genetic Algorithm based controllers improve the performance of the process in terms of time domain specification, set point tracking and disturbance rejection with optimum stability.

Keywords: PID Tuning, Genetic Algorithm, Plastic Extrusion System, Temperature Controller

Nomenclature

k	static gain
τ	apparent time delay
T	apparent time constant
L	delay time
$c(t)$	output response
K_p	proportional gain
K_i	integral gain
K_d	derivative gain
T_i	integral time
T_d	derivative time
ISE	integral square error
ITSE	integral time square error
ITAE	integral time average error
IAE	integral average error

1. Introduction

Over the years control of process system plants in the industry is customarily done by experts through the conventional PID control techniques. This is due to its simplicity, low cost design and robust performance in a wide range of operating conditions. According to an estimate nearly 90% of the controllers used in industries are PID controllers. The family of PID controllers is rigidly known as the building blocks of control theory owing to their simplicity and easy of implementation. Although the PID controllers have gained widespread usage across technological industries, it must also be pointed out that the unnecessary mathematical rigorosity, preciseness and accuracy involved with the design of controllers have been a major drawback. Designing and tuning a PID controller appears to be conceptually intuitive, but can be hard in practice, if multiple objectives are to be achieved [1]. However various techniques and modifications to the conventional PID controllers have been employed in order to overcome these difficulties. This includes the use of auto tuning PID controllers, adaptive PID controllers and also the implementation of compensation schemes to the PID controllers. But the PID controllers are with some drawback [2].

Genetic Algorithm which are adopted from the biological evaluation, are efficient search techniques that manipulate the coding representing a parameter set to search a near optimal solution through cooperation and competition among the potential solutions. These algorithms are highly relevant for industrial applications, because they are capable of handling problems with nonlinear constraints, multiple objectives and dynamic components. Genetic Algorithm is composed of two main elements which are strongly related to the problems being solved the encoding scheme and the evaluation function. The encoding scheme is used to represent the possible solutions to the problem. Individuals can be encoded in some alphabets, like binary strings, real numbers and vectors. While applying Genetic Algorithm practically, a population pool of chromosomes is installed and they are set to a random value. In each cycle of genetic



evaluation, a subsequent generation is created from the chromosomes in the current population. The cycle of evaluation is repeated until a termination criterion is reached. Fitness value can be set as the termination criterion.

In this paper Genetic Algorithm is proposed to improve the performance of a temperature control in plastic extrusion system. Tuning the controller to achieve good closed loop performance is imperative. However the effectiveness of the designed controller in terms of system requirements is greatly governed by the accuracy of the system model. Hence it is desirable to model the dynamics of the processes near the operating point by simple models such as first order process with the time delay for the purpose of controller design and also for process analysis [3]. The PID controller is designed based on traditional tuning techniques and an introduction of Genetic Algorithm is discussed in detail along with the implementation of the proposed Genetic Algorithm. The simulation results are presented and justification for the Genetic Algorithm controller is given based on time domain analysis, set point tracking and disturbance rejection. Details of plant model, PID control algorithm, proposed controller, simulation model, results and discussion, Conclusion, and References are presented in the subsequent systems.

2. Plastic Extrusion System Model

The major role of the controller is to maintaining the temperature in plastic extrusion machine as stable one. The temperature system has nonlinearity, long delay time, large time constant and undetermined system. The plastic extrusion process is a well known technique widely used in the polymerization industry. In this process the polymer is fed into the hopper in solid pellet form and it passes through the temperature zones where it is heated and melted. The melted polymer material is pushed forward by a powerful screw and it passes through the melting mechanism to form the die. An extrusion barrel in such a process usually consists of several temperature zones controlled by many electrical heaters with appropriate power specifications to provide different heating ranges [4]. In order to produce a good quality of plastic extrudates the temperature in each zone must be appropriately set and precisely controlled. The plant model of the plastic extrusion system is shown in Figure.1. The block diagram of the plastic extrusion system consists of heating stages like barrel zone, adapter, die zone is shown in Figure 2. The proposed system maintains the temperature at different set points and control sudden input disturbances in temperature change.

3. PID Control Algorithm

The system considered as mathematical model is to be analyzed as a closed loop system. In the closed loop system it is considered with unity feedback and a PID controller is placed in series. The basic representation of a PID controller is given as $C(S) = K_p + K_i/S + K_dS$, where

K_p is proportional Gain, K_i is the integral Gain, K_d is the derivative Gain. Numerous methods have been developed for setting the parameters of a PID control.



Figure 1. Plant model of a plastic extrusion system

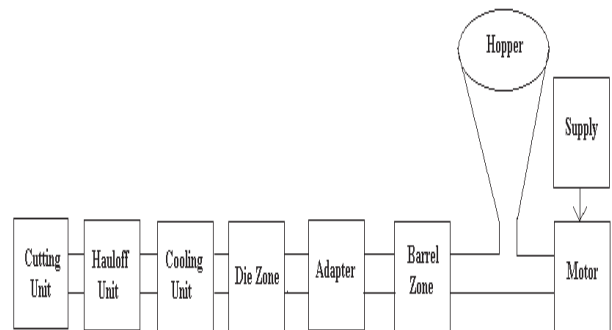


Figure 2. Block diagram of a plastic extrusion system

The PID controller settings Gain values and time constants are designed using Zeigler-Nichols tuning method [5]. By applying step test to equation (2). Step response method is based on transient response tests. Step input is applied to the transfer function of the system. Many industrial processes have step responses in which the step response is monotonous after an initial time. Closed loop control system with disturbance input is shown in Figure 3. The plastic extrusion process control system uses first order transfer function. The first order transfer function is given in (1). The plastic extrusion system transfer function is adopted from literature [6] and shown in (2). The plastic extrusion model uses the parameters $K=0.92$, $T=144S$, $\tau=10S$ seconds. S-shaped curve of temperature control model of plastic extrusion obtained as shown in Figure 4. From the S-shaped curve of step response, it is identified that the temperature control model characterized by two constants, as delay time $L=10S$ seconds and time constant $=50S$ seconds. The delay time and time constant are determined by drawing a tangent line at the inflection point of the s-shaped curve and determining the intersections of the tangent line with the time axis and line output response $C(t)$. In this paper the controller tuning is done with Zeigler-Nichols traditional tuning methodology [7] and proposed technique is with tuning method based on Genetic Algorithm.



$$G(s) = \frac{k}{1+sT} e^{-s\tau} \quad (1)$$

$$G(s) = \frac{0.92}{1+144s} e^{-10s} \quad (2)$$

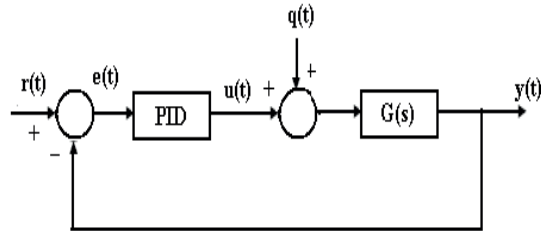


Figure 3. Closed loop control system with disturbances input

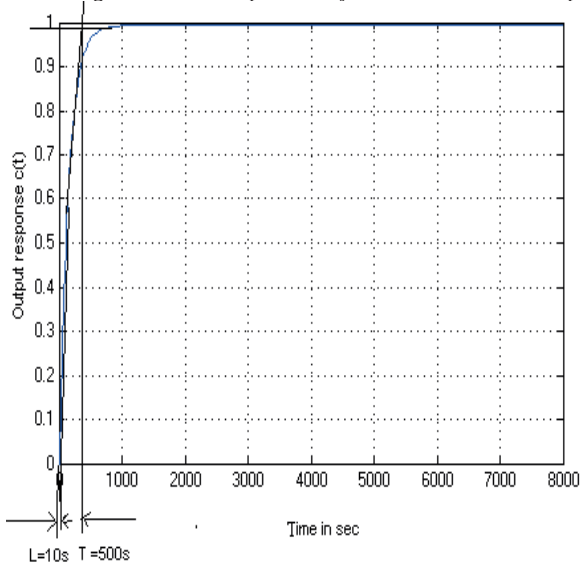


Figure 4. S shaped curve response of a temperature control model

From Zeigler-Nichols tuning rules the suggested optimal set values obtained [8]. The PID control optimal setting values for temperature control of plastic extrusion model are obtained by finding the minimum values of ISE, ITSE, ITAE, and IAE as shown in Table 1. The optimal setting values K_p , K_i , K_d obtained from the Zeigler Nicholas tuning rules and the values shown in Table 2. The simulink model of PID controller is shown in Figure 5.

Table 1. Minimum Setting Values ISE, ITSE, ITAE, IAE

ISE	ITSE	ITAE	IAE
8.054e+6	5.971e+10	1.962e+9	2.768e+5

Table 2. Minimum Setting Values K_p , K_i , K_d

K	Ti	Td	Ki	Kd
p				
30	20	05	0.315 e+10	11.52

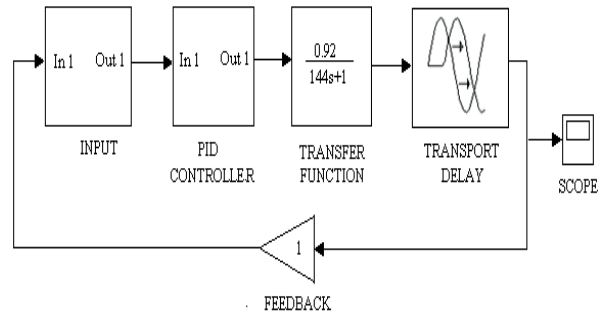


Figure 5. Simulink model of PID controller

4. Genetic Algorithm

The basic principles of Genetic Algorithm were first proposed by Holland. Genetic Algorithm was inspired by the mechanism of natural selection, a biological process in which stronger individual is likely to be the winners in a competing environment. Genetic Algorithm uses a direct analogy of such natural evolution to do global optimization in order to solve highly complex problems. It presumes that the potential solution of a problem is an individual and can be represented by a set of parameters [9]. These parameters are regarded as the genes of a chromosome and can be structured by a string of concatenated values. The form of variables representation is defined by the encoding scheme. The variables can be represented by binary, real numbers, or other forms, depending on the application data. Its range, the search space, is usually defined by the problem. Genetic Algorithm has been successfully applied to many different problems. It has also been applied to machine learning, dynamic control system using learning rules and adaptive control [10]. An illustrative flowchart of the Genetic Algorithm implementation is presented in Figure 6.

5. Genetic Algorithm Based Tuning of the Controller

The optimal value of the PID controller parameters K_p , K_i , K_d is to be found using Genetic Algorithm. All possible sets of controller parameter values are particles whose values are adjusted to minimize the objective function [11]. For the PID controller design, it is ensured the controller settings estimated results in a stable closed loop system.

Initialization of Parameters

To start with Genetic Algorithm, certain parameters need to be defined. These include population size, bit



length of chromosome, number of iterations, selection, crossover and mutation types. Selection of these parameters decides, to a great extent, the ability of the designed controller. The range of the tuning parameters is considered between 0 and 10 [12]. Initializing values are detailed as follows:

Population type: Double vector

Population size: 100

Bit length of the considered chromosome: 6

Number of generations: 100

Selection function: Tournament selection

Crossover type: Single point crossover

Crossover function: Intermediate

Crossover probability: 1.0

Mutation type: Uniform mutation

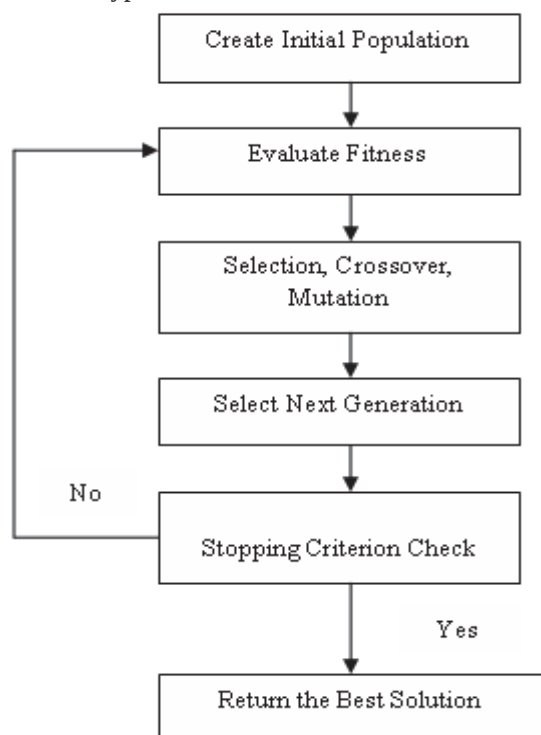


Figure 6. Flow Chart of Genetic Algorithm implementation.

In each generation, the genetic operators are applied to selected individuals from the current population in order to create a new population [13]. Generally, the three main genetic operators of reproduction, crossover and mutation are employed. By using different probabilities for applying these operators, the speed of convergence can be controlled. Crossover and mutation operators must be carefully designed, since their choice highly contributes to the performance of the whole Genetic Algorithm [14].

Reproduction

A part of the new population can be created by simply copying without change selected individuals from the present population. Also new population has the possibility of selection by already developed solutions. There are a number of other selection methods available and it is up to the user to select the appropriate one for each process. Reproduction crossover fraction is using 0.8.

Crossover

New individuals are generally created as offspring of two parents (i.e., crossover being a binary operator). One or more so called crossover points are selected (usually at random) within the chromosome of each parent, at the same place in each [15]. The parts delimited by the crossover points are then interchanged between the parents. The individuals resulting in this way are the offspring. Beyond one point and multiple point crossover, there exist some crossover types. The so called arithmetic crossover generates an offspring as a component wise linear combination of the parents in later phases of evolution it is more desirable to keep individuals intact, so it is a good idea to use an adaptively changing crossover rate: higher rates in early phases and a lower rate at the end of the Genetic Algorithm [16].

Mutation

A new individual is created by making modifications to one selected individual. The modifications can consist of changing one or more values in the representation or adding/deleting parts of the representation. In Genetic Algorithm, mutation is a source of variability and too great a mutation rate results in less efficient evolution, except in the case of particularly simple problems [17]. Hence, mutation should be used sparingly because it is a random search operator; otherwise, with high mutation rates, the algorithm will become little more than a random search. Moreover, at different stages, one may use different mutation operators. At the beginning, mutation operators resulting in bigger jumps in the search space might be preferred [18]. Later on, when the solution is close by a mutation operator leading to slighter shifts in the search space could be favored. The Figure 7. shows the simulation diagram of GA tuned PID controller.

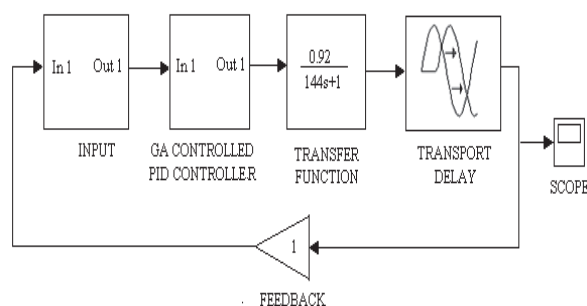


Figure 7. Simulation diagram of GA based PID controller

6. Simulation Model

The system is a multistage and coupled system so four set points is taken for the system. Four set point temperatures 70°C, 100 °C, 150 °C and 200 °C changes at different times are used in 14000 seconds. The PID simulation output of temperature control in plastic extrusion system using Matlab is shown in Figure 8. From the output of PID it is observed that the controller output is with very high initial transient overshoot, almost oscillating and take more time to settle with the reference temperature. The Genetic Algorithm based PID tuning output is shown in Figure 9. The genetic Algorithm for PID tuning has



no initial transient overshoot and oscillations and gives better results compared to PID controller. The combined results of PID (Zeigler-Nichols) and Genetic Algorithm tuning of PID is shown in Figure 10 for the temperature set point 70°C. The combined results of PID (Zeigler-Nichols) and Genetic Algorithm tuning of PID is shown in Figure 11 for the four set point temperatures.

7. Results and Discussion

The Genetic Algorithm tuned PID controller reveals shorter settling time. Moreover the overshoot is lower than the results obtained by Zeigler Nichols tuned controller. The comparisons of the performance using Genetic Algorithm based PID controller and Zeigler

Nichols are made as shown in Table.3. From the analysis the Genetic Algorithm based PID controller on the basis of delay time it gives efficient output differences 4 times to that of PID Zeigler Nichols tuning controller. Consequently the Genetic Algorithm based PID controller produces an output which is 1.67 times ahead of that of PID in the rise time analysis. The peak time results states that Genetic Algorithm based PID controller outcasts a production output times 600 efficient than PID. With consideration over the settling time the Genetic Algorithm PID controller is efficient than 1.08 times. The set point tracking and disturbance rejection is obtained in the proposed method

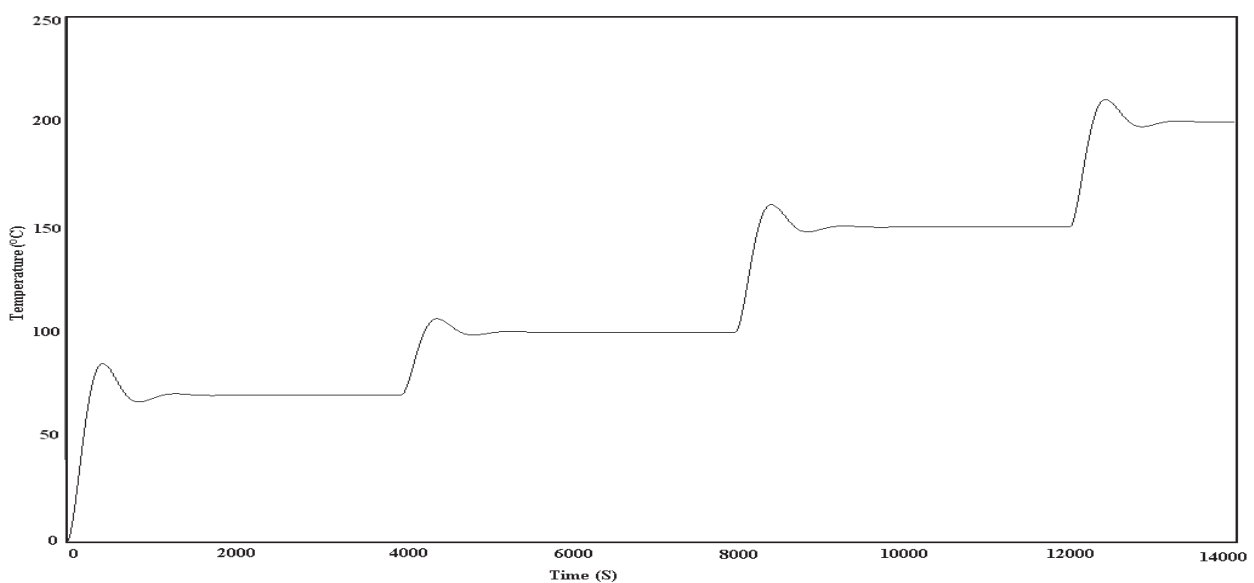


Figure 8. Simulation output of PID controller

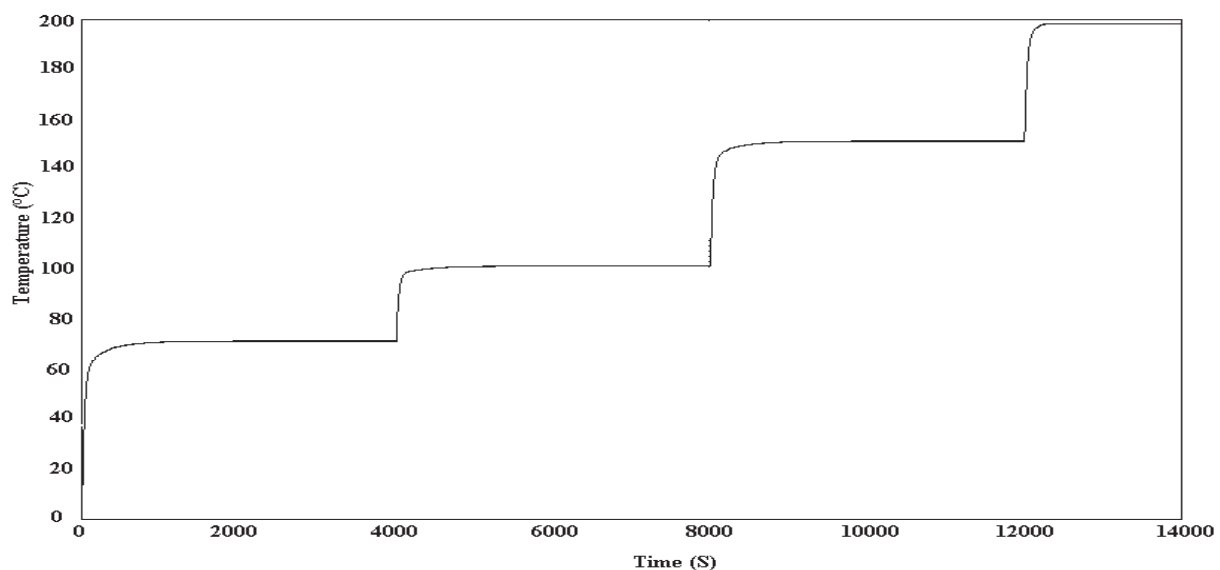


Figure 9 Simulation output of Genetic Algorithm based PID



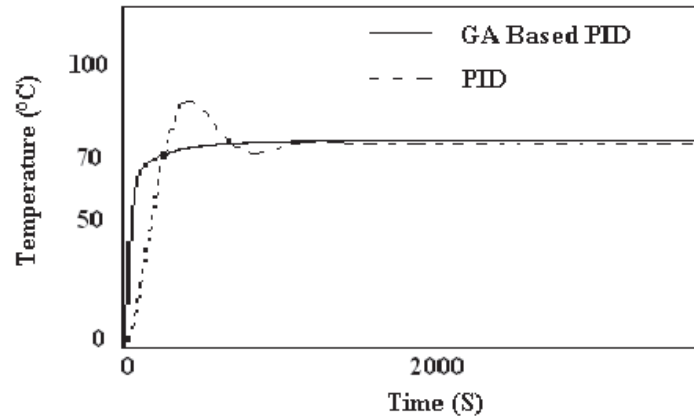


Figure 10 Combined Simulation output of Genetic Algorithm based PID and Zeigler Nichols PID for the temperature 70 °C

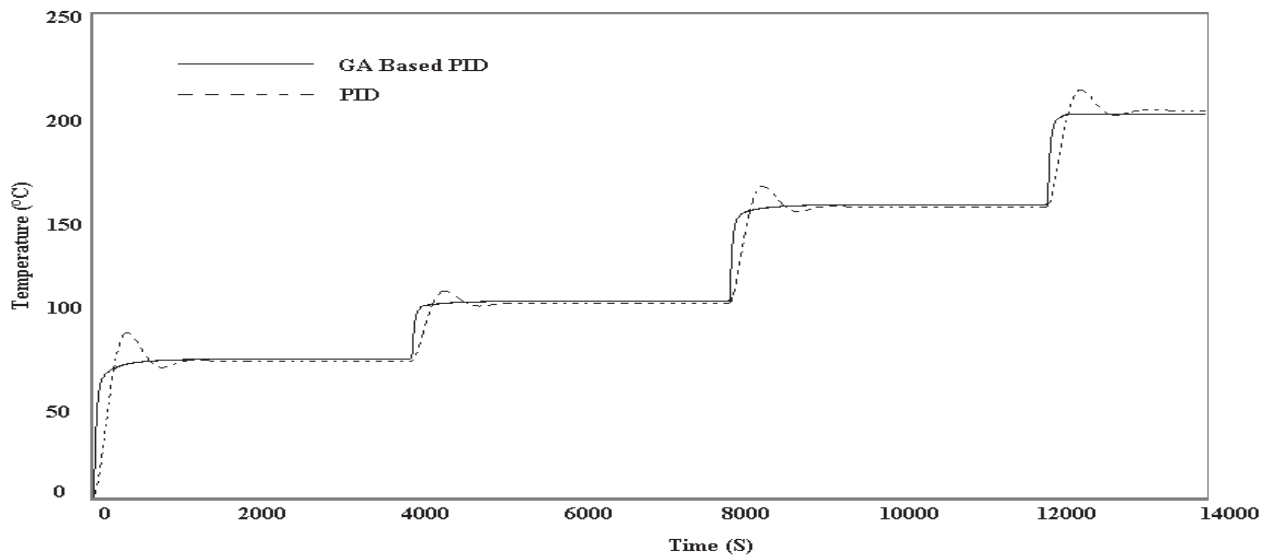


Figure 11 Combined Simulation output of Genetic Algorithm based PID and Zeigler Nichols PID

Table 3. Performance Comparison

Timing Specifications	PID	GA Tuned PID
Delay Time(t_d)	200Sec	50Sec
Rise Time(t_r)	300Sec	180Sec
Peak Time(t_p)	600 Sec	0 Sec
Settling Time(t_s)	1950 Sec	1800 Sec

8. Conclusion

This paper demonstrated the design, analysis and suitability of temperature response control model for plastic extrusion by Genetic Algorithm PID control. It uses Matlab simulink for simulation. Due to sudden input disturbances and different set points temperature changes for plastic extrusion system, the simulation result shows the PID control is with more overshoot.

The Genetic Algorithm based PID controller is without overshoot and settle with reference set point temperature level with shorter delay compared to PID controller. This paper demonstrates the effectiveness of intelligent controller on non linear system particularly for temperature control in plastic extrusion system. From the results the proposed Genetic Algorithm controller is suitable for set point changes and for stability with the aid of supervisory technique. The proposed intelligent controller identifies the process variations quickly and provides good control for the set point changes and for sudden input disturbances. The Genetic Algorithm based PID controller can adjust the PID parameters real-timely according to parameter estimation and accomplish the accurate control to complex system. The proposed system proves especially efficacious in the case of temperature control in plastic extrusion system.



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Matrix Converter Fed Induction Motor Drive Using Indirect Space Vector Modulation Technique

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Abstract

This paper presents a double-sided modulation strategy for matrix converters based upon an indirect conversion scheme which models the Matrix Converter (MC) as two independent stages performing rectification and inversion. The performance evaluation of 3Φ to 3Φ matrix converter using indirect space vector modulation (ISVM) algorithm is implemented for the Induction Motor drive. The simulation results are presented which validates the modulation technique and operation of Matrix Converter (MC).

Keywords: Matrix converter, AC-AC Converter, Indirect Space Vector modulation, Performance Evaluation, Induction Motor.

1. Introduction

A matrix converter is a direct frequency changer. This converter consists of an array of 3x3 bidirectional switches arranged so that any of the output phase of the converter can be directly connected to any input phase at any time [1].

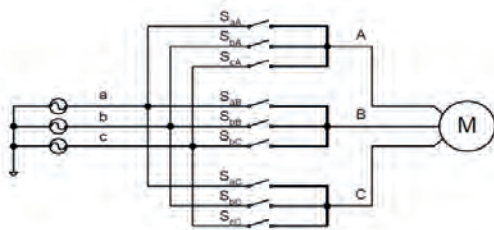


Figure 1 3-Phase Matrix Converter

In recent years, Matrix Converter (MC) has received considerable interest as a viable alternative to the conventional back-to-back PWM converter[2]-[5]. MatrixConverter reported to have attractive characteristics, i) inherent good four-quadrant operation, ii) absence of bulky dc-link electrolytic capacitors, iii) provides bidirectional power flow, iv) nearly sinusoidal I/O current waveforms and a controllable input power factor and v) increased power density.

The control of the matrix converter topology is complex. Hence various research studies were proposed to determine a suitable modulation strategy. Based on the vector approach such as the direct and indirect space vector [7] modulation (SVM and ISVM) [8, 10] methods are popular.

In this paper the ISVM strategy for three-phase matrix converter is implemented and presented. Section (2) focuses on the Fundamentals of MC space vectors and its switching states. Section (3) describes about the indirect space vector. Finally, section (4) and section (5) presents the simulation results of MC with ISVM for three phase induction motor load and conclusion respectively.

2. Fundamentals

Matrix Converter consists of an array of bidirectional switches that connects directly the load to the source. As the principle applications derived from this topology are related to three-phase electric machines, the most commonly studied matrix structure is that formed by nine bidirectional switches in a 3x3array. As far as the model is concerned, the control and modulation stage will be represented by the implementation of 9 switching functions in equation (1) for the bidirectional switches.

$$S_{Kj} = \begin{cases} 1, \text{ switch } S_{Kj} \text{ closed} \\ 0, \text{ switch } S_{Kj} \text{ open} \end{cases} \quad K = \{A, B, C\} \quad j = \{a, b, c\} \quad (1)$$

The mathematical expressions that represent the basic operation of the MC are obtained applying Kirchoff's voltage and current laws to the switch array of equations (2) and (3).

$$\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \begin{bmatrix} S_{Aa}(t) & S_{Ba}(t) & S_{Ca}(t) \\ S_{Ab}(t) & S_{Bb}(t) & S_{Cb}(t) \\ S_{Ac}(t) & S_{Bc}(t) & S_{Cc}(t) \end{bmatrix} \begin{bmatrix} v_A(t) \\ v_B(t) \\ v_C(t) \end{bmatrix} \quad (2)$$

$$\begin{bmatrix} i_A(t) \\ i_B(t) \\ i_C(t) \end{bmatrix} = \begin{bmatrix} S_{Aa}(t) & S_{Ba}(t) & S_{Ca}(t) \\ S_{Ab}(t) & S_{Bb}(t) & S_{Cb}(t) \\ S_{Ac}(t) & S_{Bc}(t) & S_{Cc}(t) \end{bmatrix}^T \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad (3)$$

The output phase voltages are v_a , v_b and v_c and i_A , i_B and i_C represents the input currents to the matrix. Since the



matrix converter is supplied by the voltage source, the input phases must not be shorted and due to the inductive nature of the load, the output phases must not be open.

$$S_{Aj} + S_{Bj} + S_{Cj} = 1, \quad j = \{a, b, c\} \quad (4)$$

The expression (4) means that, at any time, one, and only one switch must be closed in an output branch.

Table 1: Matrix Converter Switching Combinations

Group	A	b	C	V _{ab}	V _{bc}	V _{ca}	I _A	I _B	I _C	S _{Aa}	S _{Ab}	S _{Ac}	S _{Aa}	S _{Ab}	S _{Ac}	S _{Aa}	S _{Ab}	S _{Ac}
I	A	B	C	V _{AB}	V _{BC}	V _{CA}	i _a	i _b	i _c	1	0	0	0	1	0	0	0	1
	A	C	B	-V _{CA}	-V _{BC}	-V _{AB}	i _a	i _c	i _b	1	0	0	0	0	1	0	1	0
	B	A	C	-V _{AB}	-V _{CA}	-V _{BC}	i _b	i _a	i _c	0	1	0	1	0	0	0	0	1
	C	A	B	V _{CA}	V _{AB}	V _{BC}	i _c	i _a	i _b	0	0	1	1	0	0	0	1	0
	B	C	A	V _{BC}	V _{CA}	V _{AB}	i _b	i _c	i _a	0	1	0	0	1	0	1	0	0
	C	B	A	-V _{BC}	-V _{AB}	-V _{CA}	i _c	i _b	i _a	0	0	1	0	1	0	1	0	0
II A	A	C	C	-V _{CA}	0	V _{CA}	i _a	0	-i _a	1	0	0	0	0	1	0	0	1
	B	C	C	V _{AC}	0	-V _{AC}	0	i _a	-i _a	0	1	0	0	0	1	0	0	1
	B	A	A	-V _{AB}	0	V _{AB}	-i _a	i _a	0	0	1	0	1	0	0	1	0	0
	C	A	A	V _{CA}	0	-V _{CA}	-i _a	0	i _a	0	0	1	1	0	0	1	0	0
	C	B	B	V _{AC}	0	-V _{AC}	0	-i _a	i _a	0	0	1	0	1	0	0	1	0
	A	B	B	-V _{AB}	0	V _{AB}	i _a	0	-i _a	1	0	0	0	1	0	0	1	0
II B	C	A	C	V _{CA}	-V _{CA}	0	i _b	0	-i _b	0	0	1	1	0	0	0	0	1
	C	B	C	-V _{AC}	V _{AC}	0	0	i _b	-i _b	0	0	1	0	1	0	0	0	1
	A	B	A	V _{AB}	-V _{AB}	0	-i _b	i _b	0	1	0	0	0	1	0	1	0	0
	A	C	A	-V _{CA}	V _{CA}	0	0	-i _b	i _b	1	0	0	0	0	1	1	0	0
	B	C	B	V _{AC}	-V _{AC}	0	-i _b	0	i _b	0	1	0	0	0	1	0	1	0
	B	A	B	-V _{AB}	V _{AB}	0	i _b	-i _b	0	0	1	0	1	0	0	0	1	0
II C	C	C	A	0	V _{CA}	-V _{CA}	i _c	0	-i _c	0	0	1	0	0	1	1	0	0
	C	C	B	0	-V _{AC}	V _{AC}	0	i _c	-i _c	0	0	1	0	0	1	0	1	0
	A	A	B	0	V _{AB}	-V _{AB}	-i _c	i _c	0	1	0	0	1	0	0	0	1	0
	A	A	C	0	-V _{CA}	V _{CA}	-i _c	0	i _c	1	0	0	1	0	0	0	0	1
	B	B	C	0	V _{AC}	-V _{AC}	0	-i _c	i _c	0	1	0	0	1	0	0	0	1
	B	B	A	0	-V _{AB}	V _{AB}	i _c	-i _c	0	0	1	0	0	1	0	1	0	0
III	A	A	A	0	0	0	0	0	0	1	0	0	1	0	0	1	0	0
	B	B	B	0	0	0	0	0	0	0	1	0	0	1	0	0	1	0
	C	C	C	0	0	0	0	0	0	0	0	1	0	0	1	0	0	1

The Three-Phase Matrix Converter switches from Figure 1 can assume 27 allowed combinations, which are shown in Table 1. The table also shows which input and output phases are mutually connected for each allowed switching combination, as well as the resulting output line voltages and input phase currents.

3. Indirect Space Vector Modulation Algorithm

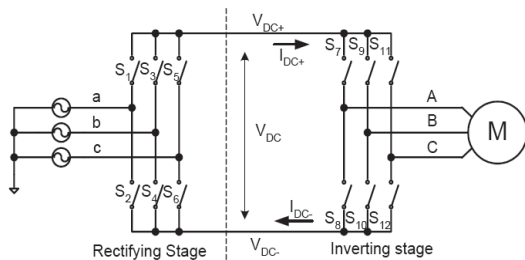


Figure 2 Emulation of VSR-VSI Conversion

The basic idea of the indirect modulation technique is to decouple the input current and the output voltage control into two stages and regards the matrix converter as a back-back PWM converter without intermediate power storage elements [4]. This is done by splitting the transfer function T of the matrix converter into the product of a

$$\begin{bmatrix} S_{aA} & S_{bA} & S_{cA} \\ S_{aB} & S_{bB} & S_{cB} \\ S_{aC} & S_{bC} & S_{cC} \end{bmatrix} = \begin{bmatrix} S_7 & S_8 \\ S_9 & S_{10} \\ S_{11} & S_{12} \end{bmatrix} \cdot \begin{bmatrix} S_1 & S_3 & S_5 \\ S_2 & S_4 & S_6 \end{bmatrix} \quad (5)$$

$$\begin{bmatrix} V_A \\ V_B \\ V_C \end{bmatrix} = \begin{bmatrix} S_7 \cdot S_1 + S_8 \cdot S_2 & S_7 \cdot S_3 + S_8 \cdot S_4 & S_7 \cdot S_5 + S_8 \cdot S_6 \\ S_9 \cdot S_1 + S_{10} \cdot S_2 & S_9 \cdot S_3 + S_{10} \cdot S_4 & S_9 \cdot S_5 + S_{10} \cdot S_6 \\ S_{11} \cdot S_1 + S_{12} \cdot S_2 & S_{11} \cdot S_3 + S_{12} \cdot S_4 & S_{11} \cdot S_5 + S_{12} \cdot S_6 \end{bmatrix} \begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \quad (6)$$

The principle of this control strategy is based on virtual DC-link in matrix converter. This DC-link is not physically present, but the switches are divided to the virtual rectifier and virtual inverter Figure 2. The indirect space vector modulation is gaining as a standard technique in the matrix converter modulation [4], [9].

Space Vector Modulation Algorithm:

In this section, based on the ITF approach, the SVM [6] is simultaneously employed for both VSR and VSI parts of the MC.

$$T = I^* R$$

VSI-SVM considers the VSI part of the circuit in Figure 2 as a standalone VSI supplied by a dc voltage source V_{dc} . The VSI switch can assume six allowed combinations which yield non zero output voltage and two



combinations with zero output voltage. These switching combinations are called switching vectors, shown in Table 2.

The expected three phase output waveform is divided into six region of each 60° . If anyone switching vector is applied, and then the inverter will give output such that, one line voltage is positive, other one is negative and third is zero.

Table 2: VSI Switching Vector

Type	Vector	$\begin{bmatrix} S_7 & S_9 & S_{11} \\ S_8 & S_{10} & S_{12} \end{bmatrix}^T$	V_A	V_B	V_C	$ V_{out} $	$\angle V_{out}$	I_{DC+}
Active	$V_1[100]$	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 1 \end{bmatrix}^T$	$2/3V_{DC}$	-	$-1/3V_{DC}$	$2/3V_{DC}$	0	I_A
	$V_2[110]$	$\begin{bmatrix} 1 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}^T$	$1/3V_{DC}$	$1/3V_{DC}$	$-2/3V_{DC}$	$2/3V_{DC}$	$\pi/3$	$-I_C$
	$V_3[010]$	$\begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 1 \end{bmatrix}^T$	$-1/3V_{DC}$	$2/3V_{DC}$	$-1/3V_{DC}$	$2/3V_{DC}$	$2\pi/3$	I_B
	$V_4[011]$	$\begin{bmatrix} 0 & 1 & 1 \\ 1 & 0 & 0 \end{bmatrix}^T$	$-2/3V_{DC}$	$1/3V_{DC}$	$1/3V_{DC}$	$2/3V_{DC}$	π	$-I_A$
	$V_5[001]$	$\begin{bmatrix} 0 & 0 & 1 \\ 1 & 1 & 0 \end{bmatrix}^T$	$-1/3V_{DC}$	-	$2/3V_{DC}$	$2/3V_{DC}$	$-2\pi/3$	I_C
	$V_6[101]$	$\begin{bmatrix} 1 & 0 & 1 \\ 0 & 1 & 0 \end{bmatrix}^T$	$1/3V_{DC}$	-	$1/3V_{DC}$	$2/3V_{DC}$	$-\pi/3$	$-I_B$
Zero	$V_0[000]$ $V_7[111]$	$\begin{bmatrix} 0 & 0 & 0 \\ 1 & 1 & 1 \end{bmatrix}^T$				0		0

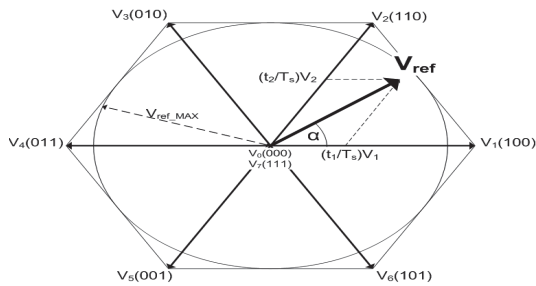


Figure 3 Inverter Voltage Hexagon

At any time any one line voltage is zero and other two becomes non-zero output. In a particular switching period, the desired voltage can be approximated by two switching vectors (V_1 and V_2) and one zero vector (V_0). Using the law of sine, duty cycles of the switching vectors are given as follows:

$$\begin{aligned} T_1 &= m_i \sin(60 - \theta_i) T_s \\ T_2 &= m_i \sin(\theta_i) T_s \\ T_{i0} &= 1 - (T_1 + T_2) \end{aligned}$$

Where m_i =modulation index of VSI,

$$m_i = (\sqrt{3} V_{ref}) / V_{DC}$$

VSR-SVM considers the VSR part of the circuit in Figure 2 as a standalone VSR. The VSR input current SVM is completely analogous to the VSI output voltage SVM. The VSI subscripts 1, 2 and i are replaced with μ, γ and r. VSR hexagon is shown in Figure 4. Using the law of sine, duty cycles of the switching vectors are given as follows:

$$\begin{aligned} T_\mu &= m_r \sin(60 - \theta_r) T_s \\ T_\gamma &= m_r \sin(\theta_r) T_s \\ T_{r0} &= 1 - (T_\mu + T_\gamma) \end{aligned}$$

Where m_r =modulation index of VSR,

$$\begin{aligned} m_r &= (\sqrt{3} V_{ref}) / V_{DC} \\ T_s &= 1/f_s \end{aligned}$$

Table 3: VSR Switching Vector

Type	Vector	$\begin{bmatrix} S_1 & S_3 & S_5 \\ S_2 & S_4 & S_6 \end{bmatrix}^T$	I_a	I_b	I_c	$ I_m $	$\angle I_m$	V_{DC}
Active	$I_1[ab]$	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}^T$	I_{DC+}	I_{DC-}	0	$2/\sqrt{3} I_{DC}$	$-\pi/6$	V_{ab}
	$I_2[ac]$	$\begin{bmatrix} 1 & 0 & 0 \\ 0 & 0 & 1 \end{bmatrix}^T$	I_{DC+}	0	I_{DC-}	$2/\sqrt{3} I_{DC}$	$\pi/6$	$-V_{ca}$
	$I_3[bc]$	$\begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}^T$	0	I_{DC+}	I_{DC-}	$2/\sqrt{3} I_{DC}$	$\pi/2$	V_{bc}
	$I_4[ba]$	$\begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \end{bmatrix}^T$	I_{DC-}	I_{DC+}	0	$2/\sqrt{3} I_{DC}$	$5\pi/6$	$-V_{ab}$
	$I_5[ca]$	$\begin{bmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \end{bmatrix}^T$	I_{DC-}	0	I_{DC+}	$2/\sqrt{3} I_{DC}$	$-5\pi/6$	V_{ca}
	$I_6[cb]$	$\begin{bmatrix} 0 & 0 & 1 \\ 0 & 1 & 0 \end{bmatrix}^T$	0	I_{DC-}	I_{DC+}	$2/\sqrt{3} I_{DC}$	$-\pi/2$	$-V_{bc}$
Zero	$I_0[aa]$ $I_7[bb][cc]$	$\begin{bmatrix} 1 & 0 & 0 \\ 1 & 0 & 0 \end{bmatrix}^T$ $\begin{bmatrix} 0 & 1 & 0 \\ 0 & 1 & 0 \end{bmatrix}^T$ $\begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & 1 \end{bmatrix}^T$				0		0

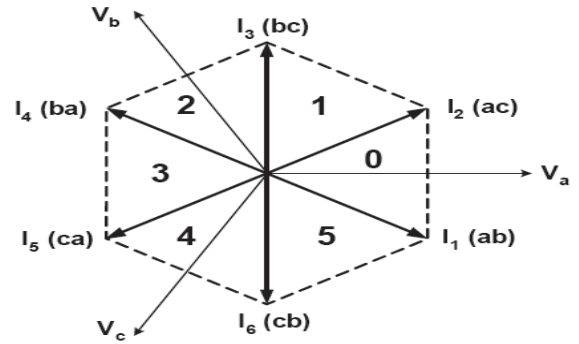


Figure 4 Rectifier Current Hexagon

Hence, VSR and VSI switching sequence are found. Using these switching sequences, MC switching sequence is calculated by using the equation (5).

4. Simulation Results

The performance of matrix converter is analyzed for closed loop three phase Induction Motor (IM). Simulations of 3HP, 220V IM with MC were made using Matlab/Simulink Software. The input filter parameters are given as $L_f=100\text{mH}$, $C_f=1\text{F}$. MC feeds induction motor with variable voltage and variable frequency of excitation power.

MC loaded by closed loop IM

Till date, much analysis is based on the assumption that the input voltages are well balanced sinusoidal and which results in the ideal output waveform. But it should be noted that harmonics would be always introduced while non-sinusoidal or unbalanced conditions are practically unavoidable. Due to lack of internal energy storage, matrix converter is highly sensitive to the disturbances in the input voltages. Therefore, it is essential to make harmonic analysis under these conditions. Estimation of harmonics in motor currents is necessary when the input voltages are non-sinusoidal. Figure 5 shows the input line current without filter and its THD is 83.56%. Figure 6



shows the input line current with filter and its THD is 52.11%. Figure 7 and Figure 8 shows the output line current without and with filter and its THD is 54.01% and 23.01% respectively.

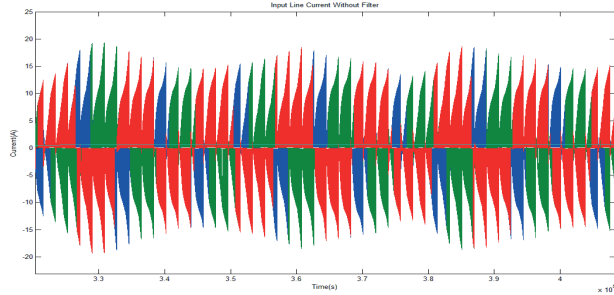


Figure 5 Input Line Current Without Filter and its THD is 83.56% at $f_o=50\text{Hz}$

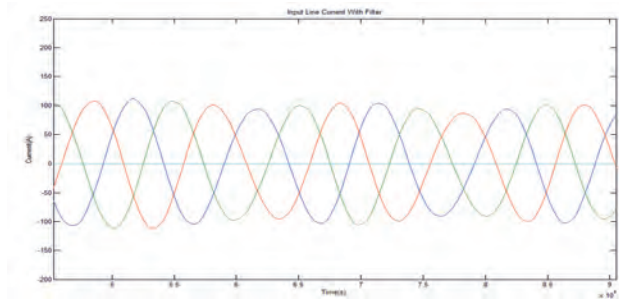


Figure 6 Input Line Current With Filter and its THD is 52.11% at $f_o=50\text{Hz}$

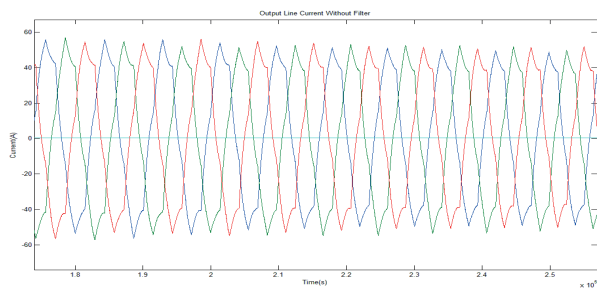


Figure 7 Output Line Current Without Filter and its THD is 54.04% at $f_o=50\text{Hz}$

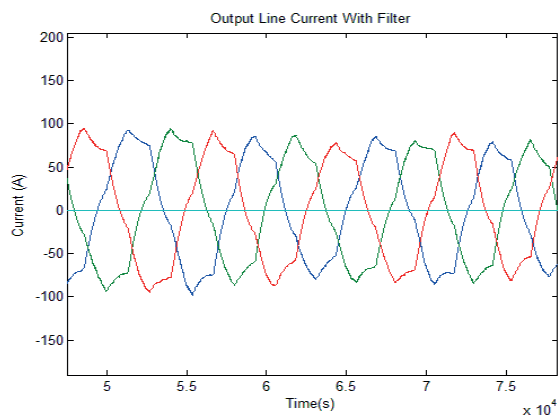


Figure 8 Output Line Current With Filter and its THD is 23.01% at $f_o=50\text{Hz}$

It is known that the supply voltage waveforms may often show typical distortion due to the presence of nonlinear load. Figure 9 shows the input voltage and current of the matrix converter without filter and the Total Harmonic Distortion of one-phase output voltage from Figure 10 is limited to 10.32%. It is seen that the input current is in

phase with the input voltage [3], which results in unity displacement factor. Figure 11 displays that three-phase output line current are essentially sinusoidal. Figure 12 and Figure 13 shows the input voltage and current of MC with filter and its THD=0.03%. Figure 14 shows the output voltage and current of MC.

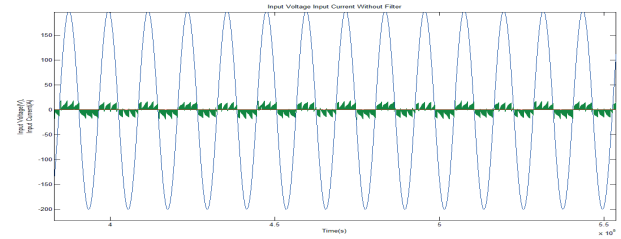


Figure 9 The matrix converter input voltage and input current without Filter

THD=10.32%

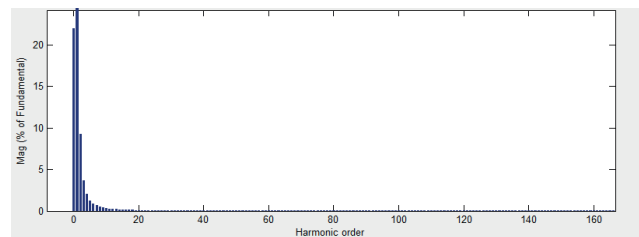


Figure 10 THD of the matrix converter input voltage and input current without Filter

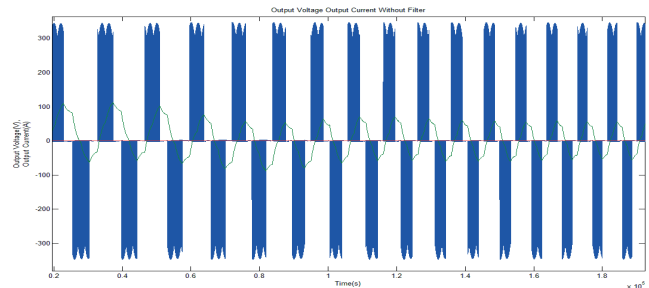


Figure 11 The matrix converter output voltage and output current without Filter

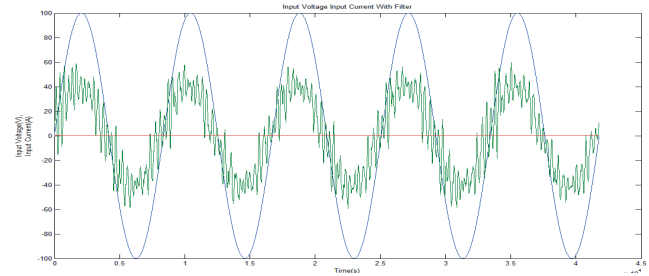


Figure 12 The Matrix C... age and input current with Filter

THD=0.03%

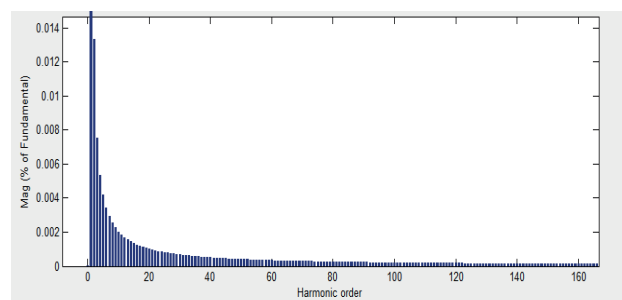


Figure 13 THD of the matrix converter input voltage and input current with Filter



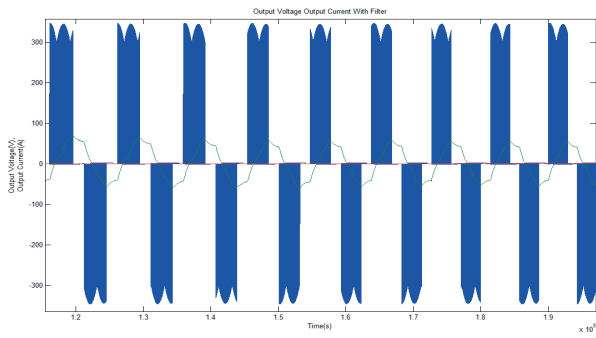


Figure 14 The matrix converter output voltage and output current with Filter

In Figure 15 the speed can reach to reference value within 0.28 seconds by no-load start. The process of speeding up is smooth and steady. Figure 16 shows the actual value of electromagnetic torque suddenly fall down to 0 Nm when actual speed reaches to reference value. Thus, induction motor dynamic performance of no-load start is excellent.

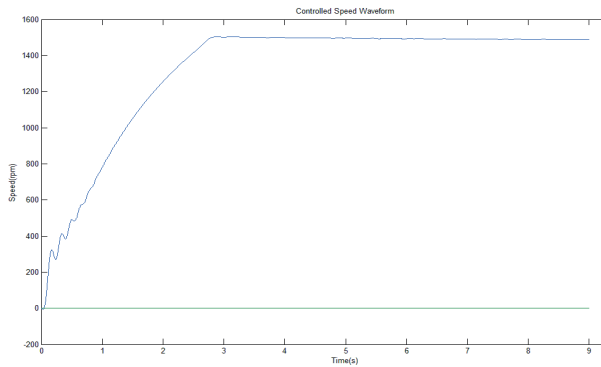


Figure 15 The Controlled Speed at 1500rpm

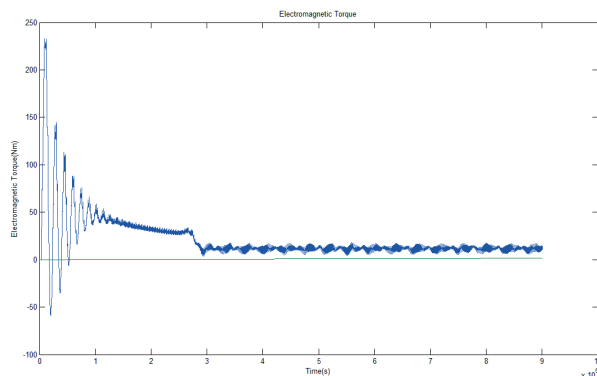


Figure 16 The Torque

5. Conclusion

Using the Indirect Conversion Scheme for a space vector modulated Matrix Converter, it is possible to model the converter as a rectifying and an inverting part, operated independent from each other. This paper, three phase to three phase matrix converter using indirect space vector modulation algorithm has been simulated for three phase induction motor load. The input filter is designed and the THD was reduced when compared to MC without Filter. The simulation results are presented through Figures 5 to 16. The inherent characteristics of MC i.e., sinusoidal input output current and unity displacement factor at the input was achieved and the controlled speed of MC using ISVM were obtained using simulation.

APPENDIX: Induction Motor Parameters

Power	2.3Kw
Supply Voltage (RMS)/Phase	220V
Frequency	50Hz
R_s	0.435 ohm
R_r	0.816 ohm
L_s	0.02H
L_r	0.02 H
L_m	0.061H
$J(\text{kg.m}^2)$	0.089
Friction Factor	0.005
Pole Pairs	2

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Genetic PD Control of a Three-Link Rigid Spatial Manipulator

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Abstract

The Lagrangian approach is employed in deriving the equations of motion for 3-link spatial manipulator carrying a payload at its end. The effects of gravity on the dynamic behavior of the manipulator are considered. The model is presented in such a way that renders the application of available control techniques an easy task.

This paper presents two schemes of robotic manipulator control systems. The first scheme is three independent PD joint controllers based on calculating the gains at the highest inertia arm configuration and; while the other scheme is based on PD controller gain parameters optimized via Genetic Algorithm (GA). Typically, for this, a controller is decomposed into a set of parameters which the GA attempts to optimize by using simulation based fitness evaluation of candidate controllers in the closed-loop system. The cost function to be minimized is the time taken by the system to achieve a final desired position. The study shows that the GA gains selection policy is effective to set the PD controller parameters in handling the rigid spatial manipulators.

Keywords: PD Control, Modeling, Rigid Manipulator, Genetic algorithms.

1. Introduction

Industrial robot manipulators are mainly positioning and handling devices. The essential problem in controlling robots is to make the manipulator follow a desired trajectory. In general, an N-degree of freedom (DOF) rigid robot manipulator is characterized by N non-linear, dynamic, coupled differential equations, [1-3]. The problem of controlling robot manipulators still offers many practical and theoretical challenges due to the complexities of the robot dynamics and the requirement to achieve high-precision trajectory tracking in the cases of high velocity movement and highly varying loads.

In the past three decades, numerous papers have been published on modeling and control of rigid robot arms, e.g., [1-6]. Generally, the Newton- Euler and Lagrangian methods are mostly used for modeling rigid robot arms. The control problem is referred to be as the motion control problem or the trajectory tracking problem. Conventional robot control methods depend heavily upon accurate mathematical modeling, analysis, and synthesis. These approaches are suitable for the control of robots that operate in structured environments. However, operations in unstructured environments require robots to perform much more complex tasks without an adequate analytical model.

Despite performance limitations, PD control is most widely used for robot position control because of its simplicity. More advanced position controllers often incorporate PD control, such as computed torque or resolved acceleration control [4]. GA's have searched to be a useful method for optimizing the PD controller gains. Genetic Algorithm (GA) based search and optimization techniques have recently found increasing use in machine learning, robot motion planning, scheduling, pattern recognition, image sensing and many other engineering applications. This paper addresses the problem of determining an optimal set of gain parameters of a PD controller for the case of a three link manipulator system. Genetic Algorithms (GAs) are search algorithms based on mechanics of natural selection and natural genetics [7]. They combine survival of the fittest among the string structures with randomized yet structured information exchange to form a search algorithm with innovative flair of natural evolution. A GA starts with a random creation of a population of strings and thereafter generates successive populations of string that improve over time [8]. The processes involved in the generation of new populations mainly consist of operations such as Reproduction, Crossover and Mutation. GAs have proven their robustness and usefulness over other search techniques because of their unique procedures



that differ from other normal search and optimization techniques in four distinct ways:

1. GAs work with coding of a parameter set, not the parameters themselves.
2. GAs search from a population of points, not a single point.
3. GAs use payoff (objective function) information, not derivative or other auxiliary knowledge.
4. GAs use probabilistic transition rules, not deterministic rules.

The present work considers the gravity effects for the more general case of spatial motion. The objective is to derive the joint angles into a desired destination starting from an arbitrary initial conditions while reducing the payload vibrations. The controller developed does not need exact dynamics of robot and is insensitive to various dynamics and payload uncertainties.

2. Dynamic Modelling of The Manipulator

The manipulator shown in Fig.1 consists of three rigid links connected by three revolute joints θ_i , $i=1,2,3$, denote the joint angles which also serves as the generalized coordinates. The mass per unit length and the length of link- i are denoted, respectively, by ρ_i and a_i . In the sequel, the following notation is adopted.

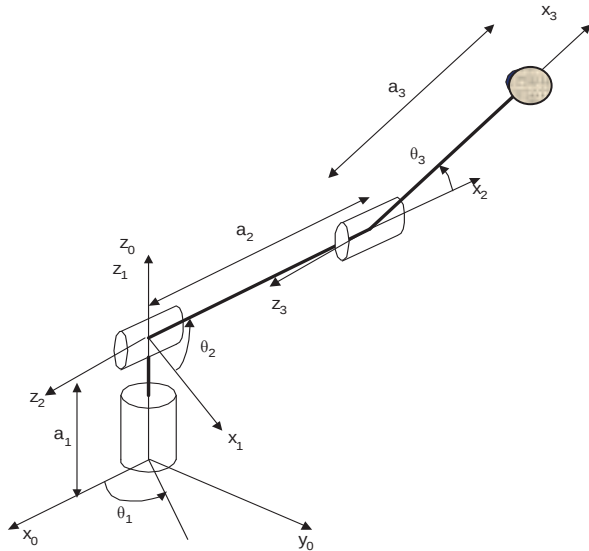


Figure 1. Three-Link Manipulator

The model developed in Appendix-A will be simulated, using MATLAB, to obtain the time response of the dynamic model. Each of the input torques at the joints is assumed to be a step function. The system dynamics equations can be rewritten in a form suitable for MatLab simulation (state-space form) as:

$$\left. \begin{aligned} \dot{y}_1 &= y_4 \\ \dot{y}_2 &= y_5 \\ \dot{y}_3 &= y_6 \\ A_{11} \dot{y}_4 &= -(C_{112} y_4 y_5 + C_{113} y_4 y_6) + u_1 \\ A_{22} \dot{y}_5 + A_{23} \dot{y}_6 &= -(C_{211} y_4^2 + C_{233} y_6^2 + C_{232} y_5 y_6) + u_2 \\ A_{32} \dot{y}_5 + A_{33} \dot{y}_6 &= -(C_{311} y_4^2 + C_{322} y_5^2 + C_{323} y_5 y_6) + u_3 \end{aligned} \right\} \quad (1)$$

Where, the state space vector y , $\dot{y}$ and u are defined as

$$y = [\theta_1 \ \theta_2 \ \theta_3 \ \dot{\theta}_1 \ \dot{\theta}_2 \ \dot{\theta}_3]^T,$$

$$\dot{y} = [\dot{y}_1, \dot{y}_2, \dot{y}_3, \dot{y}_4, \dot{y}_5, \dot{y}_6]^T \text{ and } u = [\tau_1 \ \tau_2 - D_2 \ \tau_3 - D_3]^T$$

The state model representing by Eqns.(1) can be put in the matrix form as:

$$M(t) \dot{y} = f(t, y, u) \quad (2)$$

where,

$$M(t) = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & A_{11} & 0 & 0 \\ 0 & 0 & 0 & 0 & A_{22} & A_{23} \\ 0 & 0 & 0 & 0 & A_{32} & A_{33} \end{bmatrix},$$

$$f(t, y, u) =$$

$$\begin{bmatrix} y_4 \\ y_5 \\ y_6 \\ -(C_{112} y_4 y_5 + C_{113} y_4 y_6) + u_1 \\ -(C_{211} y_4^2 + C_{233} y_6^2 + C_{232} y_5 y_6) + u_2 \\ -(C_{311} y_4^2 + C_{322} y_5^2 + C_{323} y_5 y_6) + u_3 \end{bmatrix}$$

The above equations describe the dynamic behaviour of the 3-link rigid manipulator in state space representation which is suitable for computer simulation purposes.

3. Genetic Algorithm Based Control

Recently, GA has been recognized as an effective technique to solve optimization problems and compared with other optimization techniques; GA is superior in avoiding local minima which is a common aspect in nonlinear systems [7]. GA starts with an initial population containing a number of chromosomes where each one represents a solution of the problem which performance is evaluated by a fitness function. Fig. 2 shows the flow chart of the GA process. The solution expression of the bit code is decoded to values used in an application task. This is phenotype expression. The multiple possible solutions are applied to the application task and evaluated by each. These evaluation values are fitness values. GA feed-backs the fitness values and selects current possible solutions according to their fitness values. They are parent solutions that determine the next searching points, [8]. While using GA's, the following observations are important to note:



1. GA requires no prior knowledge of the system. It is a purely data driven algorithm and is able to deal with completely unknown systems where other optimization methods may fail.
2. GA cannot be used as an on-line optimization strategy because GA cannot evaluate performance of a system at each step.

GA's consist of three basic operations: reproduction, crossover, and mutation. Reproduction is the process where members of the population reproduced according to the relative fitness of the individuals, where the chromosomes with higher fitness have higher probabilities of having more copies in the coming generation. There are a number of selection schemes available for reproduction, such as "roulette wheel," "tournament scheme," "ranking scheme," etc. [9].

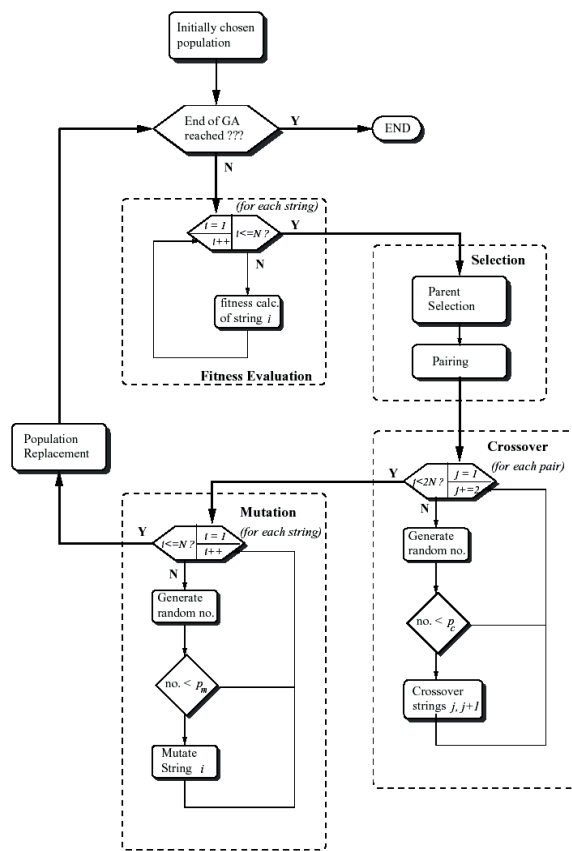


Figure 2. Flow chart of the genetic algorithm

Crossover in GA occurs when the selected chromosomes exchange partially their information of the genes, i.e., part of the string is interchanged within two selected candidates. Mutation is the occasionally alteration of states at a particular string position. Mutation is essentially needed in some cases where reproduction and crossover alone are unable to offer the global optimal solution. An example of mutation is shown in Fig.3. Further discussion on GA's can be obtained in [10-11].

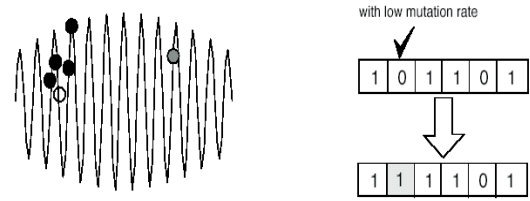


Figure 3. Effect and an example of mutation

The performance of a PD controller depends upon the proportional gain (K_p) and derivative gain (K_d). Dependent upon the values of K_p and K_d are the stability, settling time, maximum overshoot and many other system performance indicators. The proposed strategy utilizes GA as an optimization and search tool to determine the optimum value of the controller gains. There are three links and three control laws:

$$\tau_i = K_{p_i} e_i + K_{d_i} de_i/dt, \quad \text{where } i=1,2,3 \quad (3)$$

A fitness evaluation function is needed to calculate the overall responses for each of the sets of PD values and from the responses generates a fitness value for each set of individuals expressed by:

$$f(t) = \int_0^t (|e_1(t)| + |e_2(t)| + |e_3(t)|) dt \quad (4)$$

Here the goal is to find sets of PD parameters that will give a minimum fitness value over the period $[0, t]$. The GA initializes a random set of population of these six variables. The algorithm evaluates all members of the population based on the specified performance index. The GA then applies the various operations such as reproduction, crossover and mutation to generate a new set of population based on the performance of the members of the population [12]. The best member (or gene) of the population is chosen and saved for next generation. It again applies all operations and selects the best gene among the new population. The best gene of the new population is compared to best gene of previous population. If the predefined termination criterion is not met, again a new population is obtained using various operators that would have better gene. The termination criterion may be formulated as the magnitude of difference between index value of previous generation and present generation becoming less than a prespecified value. The process continues till the termination criterion is fulfilled.

4. Simulation Results

In the simulation study presented here, only set-point-tracking is considered. The desired values for the twist, shoulder, and elbow joint angles are $\theta_d = 10 \text{ deg}$, $\theta_{2d} = 15 \text{ deg}$, and $\theta_{3d} = 20 \text{ deg}$. The numerical example considered is assumed to have the following physical properties for the links $a_1 = 0.35 \text{ m}$, $a_2 = 0.85 \text{ m}$, $a_3 = 0.52 \text{ m}$, $m_T = 0.50 \text{ kg}$, $j_1 = 0.10 \text{ kg m}^2$, $\rho_2 = 2 \text{ kg/m}$, $\rho_3 = 2 \text{ kg/m}$. Proportional and derivative control gains are calculated at each joint to give physically



reasonable control torque inputs. The links inertia J changes with the arm configuration, so its maximum value should be used for calculating the controller gains corresponding to the joint axis [13]. From dynamic equations one can get the equivalent moment of inertia for maximum links inertia configuration about each joint axis-1,2, and 3 respectively as:

$$\begin{aligned} J_{eq1} &= J_1 + \frac{1}{3} \rho_2 a_2^3 + \frac{1}{3} \rho_3 a_3 (a_3^2 + 3a_2 a_3 + 3a_2^2) \\ &+ m_T (a_2 + a_3)^2 \\ &= 2.6515 \text{ kg.m}^2 \\ J_{eq2} &= \frac{1}{3} \rho_2 a_2^3 + \frac{1}{3} \rho_3 a_3 (a_3^2 + 3a_2^2 + 3a_2 a_3) \\ &+ m_T (a_2^2 + a_3^2 + 2a_2 a_3) \\ &= 2.5515 \text{ kg.m}^2 \quad J_{eq3} = \frac{1}{3} \rho a_3^3 + m_T a_3^2 \\ &= 0.2083 \text{ kg.m}^2 \end{aligned}$$

$$\text{One can easily get the } J_{eq} \ddot{\theta}_1 = u_1 \quad (5)$$

where, J_{eq} is the equivalent moment of inertia about the axis of link-1. The overall robot arm dynamics become: $J_{eq} \ddot{\theta}_1 = G_{D1} \dot{e}_1 + G_{p1} e_1$

The closed-loop error dynamics are

$$\ddot{e}_1 + \frac{G_{D1}}{J_{eq}} \dot{e}_1 + \frac{G_{p1}}{J_{eq}} e_1 = 0 \quad (6)$$

where as before $e_1 = [\theta_{1f} - \theta_1]$. The standard form for the second order differential equation is

$$\ddot{\theta}_1 + 2\zeta\omega_n \dot{\theta}_1 + \omega_n^2 \theta_1 = 0 \quad (7)$$

with ζ the damping ratio and ω_n the natural frequency. Comparing the eq.(8) with eqn(3.21) one can easily get $\frac{G_{p1}}{J_{eq}} = \omega_n^2$ and $\frac{G_{D1}}{J_{eq}} = 2\zeta\omega_n$.

It is undesirable for the robot to exhibit any overshoot since this could cause impact if, for instance, a desired trajectory terminate at the surface of the work-piece. Therefore, the control gains are usually selected for critical damping where the damping ratio $\zeta = 1$. For this case the gains are chosen as:

$$G_{p1} = J_{eq1} \omega_n^2 \quad (8)$$

$$G_{D1} = 2 J_{eq1} \omega_n \quad (9)$$

In the same way, one can select the other controller gains as:

$$G_{p2} = J_{eq2} \omega_n^2, \quad G_{p3} = J_{eq3} \omega_n^2 \quad (10)$$

$$G_{D2} = 2 J_{eq2} \omega_n, \quad G_{D3} = 2 J_{eq3} \omega_n \quad (11)$$

Therefore, the proportional and derivative gains can be easily obtained if one knows the natural frequency of the system. The natural frequency ω_n governs the speed of response in each error components. It should be large for fast responses and is selected depending on the performance objectives. Thus the desired trajectories should be taken into account in selecting ω_n . There are some upper limits on the choice of ω_n , [5]. It must be

taken into consideration that the control torque produced does not reach the upper limit of actuator saturation. If the control gains are too large, the torque $u(t)$ may reach its upper limits, ω_n should be selected as $\omega_n < 0.5 \omega_r$, where ω_r is the frequency of the first resonant mode of the link in calculations. A reasonable value for ω_n is selected as 10 rad s^{-1} . The gains are then calculated using eqs. (10, 11) and are shown in Table 1.

Table 1. Controller Gains

G_{p1}	G_{D1}	G_{p2}	G_{D2}	G_{p3}	G_{D3}
265.15	53.03	255.15	51.03	20.83	4.16

The Matlab Package has been used to model the manipulator by numerically simulating the previous equations. The differential equations have been integrated by the 4th order Rung-Kutta method. The sampling time interval is taken as 0.05 s. The simulation presented here is achieved by applying step input for the joint angles and the control objective is to track the joint angles to their set values. The suggested PD-controller has the following structure

$$\tau_1 = -G_{p1}(\theta_1 - \theta_{1f}) - G_{D1}\dot{\theta}_1 = G_{p1}e_1 - G_{D1}\dot{\theta}_1 \quad (12)$$

$$\begin{aligned} \tau_2 &= -G_{p2}(\theta_2 - \theta_{2f}) - G_{D2}\dot{\theta}_2 + d(2) \\ &= G_{p2}e_2 - G_{D2}\dot{\theta}_2 + d_2 \end{aligned} \quad (13)$$

$$\begin{aligned} \tau_3 &= -G_{p3}(\theta_3 - \theta_{3f}) - G_{D3}\dot{\theta}_3 + d(3) \\ &= G_{p3}e_3 - G_{D3}\dot{\theta}_3 + d_3 \end{aligned} \quad (14)$$

where θ_{if} and $e_i = (\theta_{if} - \theta_i)$, $i=1, 2, 3$, denote the desired final joint angle and the tracking error for each joint, respectively and d_2 and d_3 are the gravity compensation terms for the links 2 and 3. Gravity compensation is not constant, but rather depends on the instantaneous link configuration. This is required to ensure global stability of the manipulator since the gravity is exactly cancelled by feedback [13,14]. The proportional and derivative gains, G_{p_i} and G_{D_i} , used in this study are chosen as in Table 2. It is convenient to write this control law in a compact form as,

$$\tau = G_p e - G_D \dot{e} + d \quad (15)$$

where $\tau = [\tau_1, \tau_2, \tau_3]^T$, $G_p = [G_{p1}, G_{p2}, G_{p3}]^T$,

$G_D = [G_{D1}, G_{D2}, G_{D3}]^T$, and $d = [d_1, d_2, d_3]^T$

Table 2. Control gains

	Joint 1	Joint 2	Joint 3
G_{p_i}	797	701	198
G_{D_i}	39.6	30.4	4.96

These values of gain parameters were used to simulate system response for a run time of 2 seconds. Figs. 4 and 5 show the joint angles step response and the corresponding joint velocities plotted versus time, while Fig. 6 show the applied joints torque plotted against time. It shows a settling time of approximately from 0.6s to 1 seconds for the three arms. It is seen that the controller performance



characterized by an overshoots for all the angles. Increasing the derivative gains slowly by performing some simulations until a satisfactory performance is achieved can eliminate this small overshoot but it will decrease the speed of response as shown in Fig. 7.

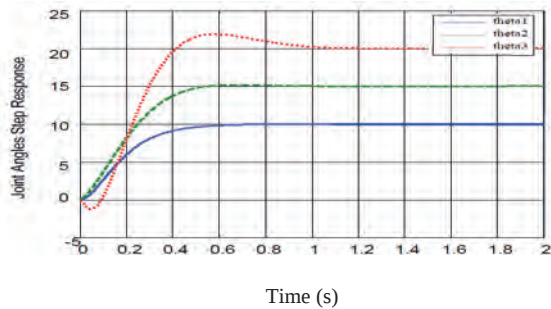


Figure 4 Joint angle step response based on PD control

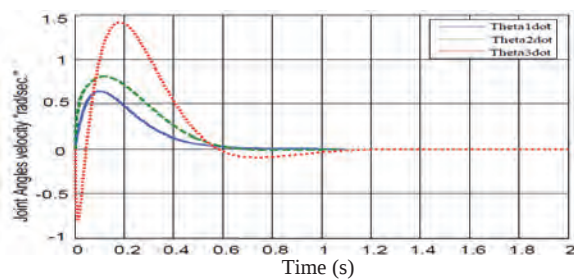


Figure 5 Joint angle velocity based on PD control

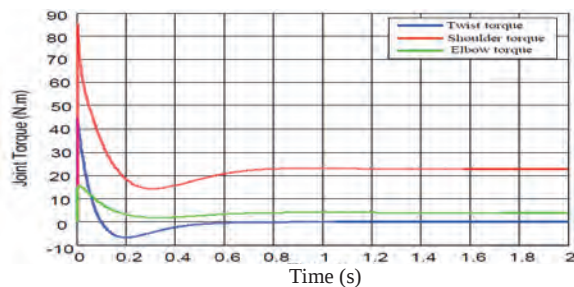


Figure 6 Joint Torque based on PD control

The GA uses its operators and functions to find the values of K_{p1} , K_{d1} , K_{p2} , K_{d2} , K_{p3} , and K_{d3} for which the performance index is a minimum. These values of gains lead to the optimum values of gains for which the system attains final position in minimum time. For each of these parameters, a population of 100 elements has been created to find the optimum parameters. Figures 8 and 9 show the trip of the proportional and derivative gains during the optimization process via GA. It is noticed that trip takes about 20 generations to arrive its optimal values as illustrated in Fig. 10.

The optimum values of parameters obtained are: proportional Gain values are $k_{1p}=751.55$; $k_{2p}=278.12$; $k_{3p}=90.39$; and Derivatives gain values are $k_{1d}=37.66$; $k_{2d}=23.89$; $k_{3d}=9.36$. These values of gain parameters were used to simulate system response for a run time of 2 seconds.

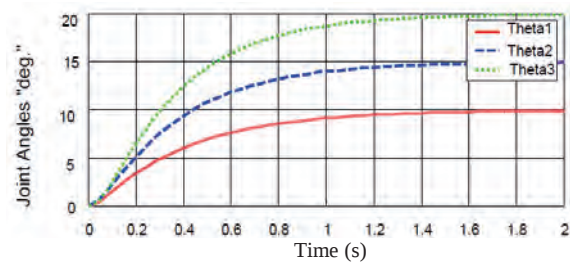


Figure 7 Joint angle step response based on PD control

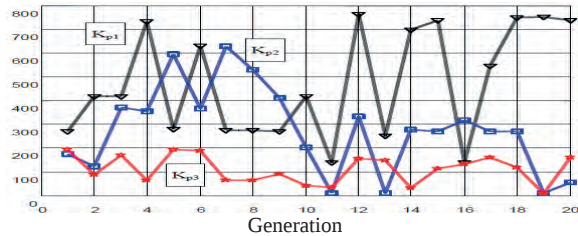


Figure 8 K_{p1} , K_{p2} and K_{p3} gains during the optimization process

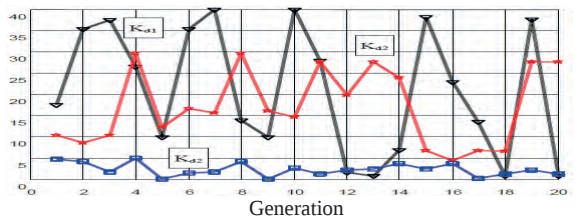


Figure 9 K_{d1} , K_{d2} and K_{d3} gains during the optimization process

Figure 11 shows angles made by both the joints plotted against time for GA gains controller based design. It also shows a settling time of approximately 0.4 seconds which is smaller when compared to PD controller. Moreover, GA based controller doesn't show any overshoot while PD controller shows overshoot in angle made by first arm. Figs 12, 13 illustrate the joints velocities and their applied torques.

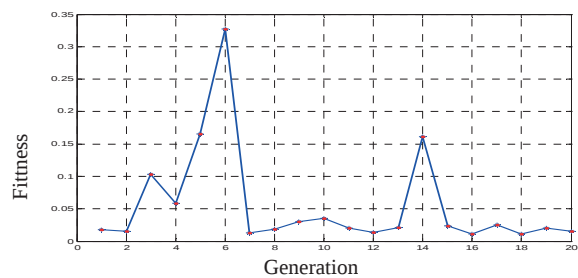


Figure 10 Fitness values with respect to the number of generation

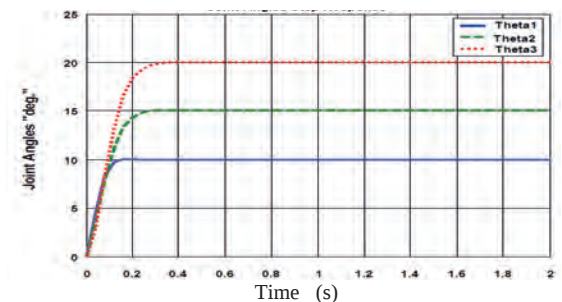


Figure 11 Joint angle step response based on GA control



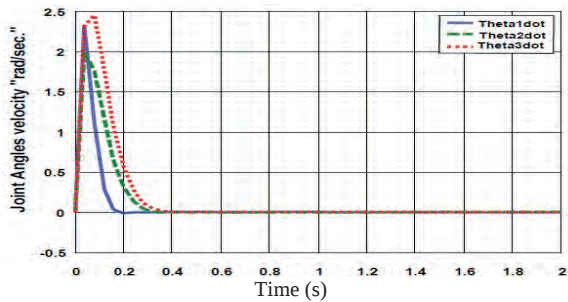


Figure 12 Joint angle velocity based on GA control

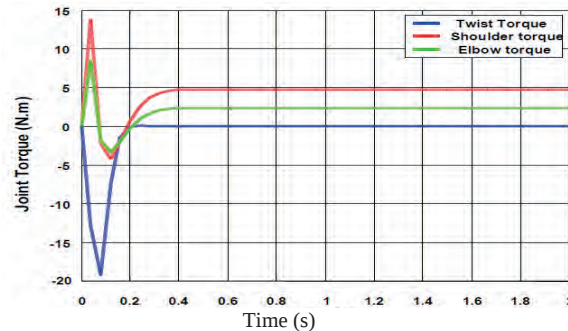


Figure 13 Joint Torque based on GA control

5. Conclusions

A PD controller has been designed in the present study and its gains have been optimized with the help of Genetic Algorithm to yield minimum convergence time. The results have been compared with a typical PD controller. The two control strategies have been implemented on a three link manipulator system. Due to the strong nonlinear characteristics and parameter variations in real environments, tracking control of a robot arm system is difficult. Genetic PD controller performs well compared to PD controllers, but only problem faced with these controllers is overcome by tuning of scaling factors with Genetic algorithm as they do not rely heavily on the designer but they involve some processing time. Future work in a related field could include not only optimizing the controller parameters but also optimizing the controller structure with the help of GA.

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Appendix A: Dynamic modelling of 3-link manipulator

The equations of motion of the manipulator derived below by using the Lagrangian approach. Toward that end, the kinetic energy and potential energy expressions of each link as well as the payload will first be obtained.

$$s_i = \sin \theta_i, \quad s_{ij} = \sin(\theta_i + \theta_j), \quad s_{ij} = \sin(2\theta_i + \theta_j) \\ c_i = \cos \theta_i, \quad c_{ij} = \cos(\theta_i + \theta_j), \quad c_{ij} = \cos(2\theta_i + \theta_j)$$

Energy Expression for link-1 the kinetic and potential energies of link- 1 are easily obtained as:

$$KE_1 = \frac{1}{2} J_1 \dot{\theta}_1^2, \quad PE_1 = 0 \quad (A-1)$$

Energy expression for link-2

$$KE_2 = \frac{1}{6} \rho_2 a_2^3 (\dot{\theta}_2^2 + \dot{\theta}_1^2 c_2^2), \quad PE_2 = \frac{1}{2} \rho_2 g a_2^2 s_2 \quad (A-2)$$

Energy Expression for link- 3

The kinetic energy of link-3 can be obtained as:

$$KE_3 = \frac{1}{2} \dot{\theta}^T M \dot{\theta}, \quad \text{Where, mass matrix } M \text{ for link-3}$$

$$\begin{bmatrix} m_{11} & m_{12} & m_{13} \\ m_{21} & m_{22} & m_{23} \\ m_{31} & m_{32} & m_{33} \end{bmatrix}$$

has the form where its elements which represent the inertia mass coefficients of link-3, are derived as:

$$m_{11} = \frac{1}{3} \rho_3 a_3^3 c_{23}^2 + \rho_3 a_2 a_3^2 c_2 c_{23} + \rho_3 a_2^2 a_3 c_2^2;$$

$$m_{22} = \frac{1}{3} \rho_3 a_3^3 + \rho_3 a_2 a_3^2 c_3 + \rho_3 a_2^2 a_3;$$

$$m_{33} = \frac{1}{3} \rho_3 a_3^3;$$



$$m_{23} = m_{32} = \frac{1}{3} \rho_3 a_3^3 + \frac{1}{2} \rho_3 c_3 a_2 a_3^2 ;$$

$$m_{12} = m_{21} = m_{13} = m_{31} = 0 .$$

Thus, The kinetic energy and potential energy of link-3 is finally obtained as:

$$KE_3 = \frac{1}{2} (m_{11} \dot{\theta}_1^2 + m_{22} \dot{\theta}_2^2 + m_{33} \dot{\theta}_3^2 + 2 m_{23} \dot{\theta}_2 \dot{\theta}_3)$$

$$; \quad PE_3 = \rho_3 g \left(a_3 a_2 s_2 + \frac{1}{2} a_3^2 s_{23} \right)$$

(A-3)

Energy Expression for the Payload

The payload kinetic energy is given by

$$KE_T = \frac{1}{2} m_T [(a_2 c_2 + a_3 c_{23})^2 \dot{\theta}_1^2 + (a_2^2 + a_3^2 + 2 a_2 a_3 c_3) \dot{\theta}_2^2$$

$$+ 2 a_3 (a_3 + a_2 c_3) \dot{\theta}_2 \dot{\theta}_3 + a_3^2 \dot{\theta}_3^2]$$

$$PE_T = m_T g (a_2 s_2 + a_3 s_{23})$$

(A-4)

Manipulator Equations of Motion

The general equations of motion for the manipulator based on Lagrangian dynamics can be obtained from,

$$\frac{d}{dt} \left(\frac{\partial KE}{\partial \dot{q}_i} \right) - \frac{\partial KE}{\partial q_i} + \frac{\partial PE}{\partial q_i} = \tau_i$$

where, $q_i = \theta_1, \theta_2, \text{ or } \theta_3$.

(A-5)

It is obvious that the system energy contributions come from the energies of Links 1, 2, and 3, as well as the payload at the tip of the arm. Thus, we can write the total system kinetic energy KE_s and the total system Potential energy PE_s as:

$$KE_s = KE_1 + KE_2 + KE_3 + KE_T, \quad PE_s = PE_2 + PE_3 + PE_T$$

(A-6)

The equations of motion for the three degrees of freedom θ_1, θ_2 and θ_3 , are derived as given, respectively, by

$$\left. \begin{aligned} A_{11} \ddot{\theta}_1 + C_{112} \dot{\theta}_1 \dot{\theta}_2 + C_{113} \dot{\theta}_1 \dot{\theta}_3 &= \tau_1 \\ A_{22} \ddot{\theta}_2 + A_{23} \ddot{\theta}_3 + C_{211} \dot{\theta}_1^2 + C_{233} \dot{\theta}_3^2 + C_{232} \dot{\theta}_3 \dot{\theta}_2 + D_2 &= \tau_2 \\ A_{32} \ddot{\theta}_2 + A_{33} \ddot{\theta}_3 + C_{311} \dot{\theta}_1^2 + C_{322} \dot{\theta}_2^2 + D_3 &= \tau_3 \end{aligned} \right\}$$

(A-7)

Where, τ_i ; The external torques acting at the robot joints, $i=1, 2, 3$.

The dynamic coefficients of equations (A-7) are given as follows:

$$A_{11} = J_1 + \frac{1}{3} \rho_2 a_2^3 c_2^2 + \frac{1}{3} \rho_3 a_3 (a_3^2 c_{23}^2 + 3 a_2 a_3 c_2 c_{23} + 3 a_2^2 c_2^2)$$

$$+ m_T (a_2 c_2 + a_3 c_{23})^2$$

$$A_{22} = \frac{1}{3} \rho_2 a_2^3 + \frac{1}{3} \rho_3 a_3 (a_3^2 + 3 a_2^2 + 3 a_2 a_3 c_3)$$

$$+ m_T (a_2^2 + a_3^2 + 2 a_2 a_3 c_3)$$

$$A_{33} = \frac{1}{3} \rho_3 a_3^3 + m_T a_3^2$$

$$A_{23} = A_{32} = \frac{1}{3} \rho_3 a_3 (a_3^2 + 1.5 a_3 a_2 c_3)$$

$$+ m_T a_3 (a_3 + a_2 c_3)$$

$$C_{211} = \left[\frac{1}{3} \mu_2 a_2^3 c_2 s_2 + \frac{\rho_3 a_3}{6} (2 a_3^2 c_{23} s_{23} + 3 a_2 a_3 s_{223} + 6 a_2^2 s_2 c_2) \right.$$

$$\left. + m_T (a_2 c_2 + a_3 c_{23}) (a_2 s_2 + a_3 s_{23}) \right]$$

$$C_{233} = - (0.5 \rho_3 a_3^2 a_2 s_3 + m_T a_3 a_2 s_3)$$

$$C_{112} = \frac{2}{3} \rho_2 a_2^3 c_2 s_2 - \frac{1}{3} \rho_3 a_3 (2 a_3^2 c_{23} s_{23} + 3 a_2 a_3 s_{223} + 6 a_2^2 c_2 s_2)$$

$$2 m_T (a_2 c_2 + a_3 c_{23}) (a_2 s_2 + a_3 s_{23})$$

$$C_{113} = - \frac{\rho_3 a_3}{3} (2 a_3^2 c_{23} s_{23} + 3 a_2 a_3 c_2 s_{23})$$

$$- 2 m_T (a_2 c_2 + a_3 c_{23}) a_3 s_{23}$$

$$C_{232} = - (\rho_3 a_2 a_3^2 s_3 + 2 m_T a_2 a_3 s_3)$$

$$C_{311} = \left[\frac{1}{6} \rho_3 a_3 (2 a_3^2 c_{23} s_{23} + 3 a_2 a_3 c_2 s_{23}) \right.$$

$$\left. + m_T (a_2 c_2 + a_3 c_{23}) a_3 s_{23} \right]$$

$$C_{322} = \left(\frac{1}{2} \rho_3 a_2 a_3^2 s_3 + m_T a_2 a_3 s_3 \right)$$

$$D_2 = \frac{1}{2} \rho_2 g a_2^2 c_2 + \frac{1}{2} \rho_3 g a_3 (2 a_2 c_2 + a_3 c_{23})$$

$$+ m_T g (a_2 c_2 + a_3 c_{23})$$

$$D_3 = \frac{1}{2} \rho_3 g a_3^2 c_{23} + m_T g a_3 c_{23}$$

Where,

A_{ii} : The effective inertia coefficient at joint i .

A_{ij} : The coupling inertia coefficient between joints i and j , at joint i

C_{ijj} : The centripetal force dynamic coefficient at joint i due to velocity at joint j

C_{ijk} : The Coriolis force dynamic coefficient at joint i due to velocities at joints j and k

D_i : The gravity loading at joint i .

Nonlinear, coupled, and second ordered differential equations. Eqns. (A-7) can finally be rewritten in matrix form as follows:

$$A(q) \ddot{q} + C(q, \dot{q}) + D(q) = \tau$$

(A-8)

where,

$A(q)$; is the system Inertia Matrix with elements,

$$A(q) = \begin{bmatrix} A_{11} & 0 & 0 \\ 0 & A_{22} & A_{23} \\ 0 & A_{32} & A_{33} \end{bmatrix} ;$$

$C(q, \dot{q})$; is The Coriolis /Centripetal Vector,

$$C(q, \dot{q}) = \begin{bmatrix} c_{112} \dot{\theta}_1 \dot{\theta}_2 + c_{113} \dot{\theta}_1 \dot{\theta}_3 \\ c_{211} \dot{\theta}_1^2 + c_{233} \dot{\theta}_3^2 + c_{232} \dot{\theta}_2 \dot{\theta}_3 \\ c_{311} \dot{\theta}_1^2 + c_{322} \dot{\theta}_2^2 + c_{323} \dot{\theta}_2 \dot{\theta}_3 \end{bmatrix} ;$$

$D(q)$; is the Gravity vector of the form

$$D(q) = [0 \quad D_2 \quad D_3]^T ;$$

τ ; is the external torques applied at the arm joints

$$\tau = [\tau_1 \quad \tau_2 \quad \tau_3]^T .$$





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Improvement of Transient Stability by FACTS devices

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Abstract

The rate of dissipation of transient energy is usually used as a tool to measure dynamic system damping. For such a given system equipped with flexible alternative current transmission devices, the above concept is then applied to determine the additional damping. This paper applies and discusses this control strategy for suppressing undesirable electromechanical oscillations in power system with both static synchronous series compensator and static synchronous compensators. The same devices are studied by applying a new method of injection model. Dynamic model of these devices are derived for the two control strategy. Several disturbances are studied for the single and multimachine power systems. It is found that the found results convey eminent efficiency of the described theories. MATLAB/SIMULINK software is used for all simulations.

Keywords: SSSC, STATCOM, transient energy function, damping.

Nomenclature

δ Generator rotor angle
 ω Generator rotor speed
 δ_s Rotor angle at the post-fault stable equilibrium point
 δ_m Angle of the intermediate bus m
 θ Angle of line current
 α_s Angle of the SSSC voltage
 P_e Electrical power output in per unit.
 P_m Mechanical power input in per unit.
 D Generator damping coefficient
 H Inertia constant in sec
 M Moment of inertia
 V_t Generator terminal voltage
 V_b Infinite bus voltage
 E Magnitude of the internal voltage

V_m Voltage magnitude of the intermediate bus m
 x_t Transformer reactance
 x'_d Generator transient reactance
 X_{L1} Reactance of line 1
 X_{L2} Reactance of line 2
 X_1 Equivalent reactance between machine internal bus and bus m
 X_2 Equivalent reactance between bus m and infinite bus
 X_s Reactance of the series transformer
 X_L Line reactance
 V_p Magnitude of STATCOM injected voltage
 α_p Angle of the STATCOM injected voltage
 Y_p Magnitude of the shunt admittance of STATCOM
 θ_p Argument of the shunt admittance of STATCOM
 E_i Voltage magnitude of the i-th machine behind the transient reactance.
 δ_i Rotor angle of the i-th machine behind the transient reactance.
 M_i Inertia moment of the i-th machine.
 D_i Generator damping coefficient of the i-th machine.
 P_{mi} Mechanical input power of the i-th machine.
 P_{ei} Electrical output power of the i-th machine.
 Y_{ij} Magnitude of the ij-th element of the admittance matrix reduced to the n generator of network.
 θ_{ij} Argument of the ij-th element of the admittance matrix reduced to the n generator of network.

1. Introduction

One of the major problems in an electrical power system is network stability. Several disturbances such as short



circuits, step change in the mechanical input power can lead to disequilibrium between mechanical power and electrical power. This later can affect rotor speed variations and can lead to a loss of synchronism and thus to a total system instability. After clearing perturbation network is then stable and can satisfy balance offers request of reactive energy. For such a system, the capability to maintain it's stable equilibrium point due to a sudden disturbance is an interest concern. During perturbation a transient regime is instituted presenting undesirable mechanical oscillations. In some cases, The power system stabilizer has been used as a tool to assess transient stability and suppress mechanical oscillations , but no, the development of recent FACTS (Flexible Alternative Current Transmissions Systems), such as: STATCOMs (static compensators), SSSC (static synchronous series compensator), SVC (static var compensator), and TCSC (thyristor controller switch compensator) allow a fast and effective control of damping mechanical oscillations by introducing an additional damping coefficient on the power system . The influence of these devices is well known, in literature, References [1-2] study the stability enhancement problem of a power system considering multiple STATCOMs by developing and testing fuzzy algorithms and adaptive control strategy.

Reference [3] studies the effect of SVC to improve voltage and critical clearing time.

Reference [4] apply artificial neural network (ANN) based phase shifting transformer (PST) to enhance multi-machine power system transient stability.

The lyapunov method based on the energy function is a method to assess the degree of stability [5]. This concept is always used to damp oscillations in power systems.

References [6-7-8] studies the same concept with new additional damping terms related to FACTS devices for a classical single machine infinite bus and the 10- machine New England system.

References [9-10] studies effect of FACTS devices on improving network voltage plan and adopt approach of injection model. But works are limited to static cases and neglect dynamic performances of FACTS with this approach.

This paper adopts the method to study transient stability of an electrical power system under several perturbations by using two types of FACTS devices and implement the approach of injection model to demonstrate its effect on multimachine power system stability.

The strategies are tested for the kundur's single machine infinite bus (SMIB) and a multimachine power system.

The remainder of the paper is organized as follows: Section (2) focuses on providing the basic dynamic model of a SMIB equipped with SSSC and STATCOM, Section (3) presents theoretical models of these two devices with the strategy of injection model Section (4) emphasizes an overview of simulation results.

2. Background

This section provides the basic dynamic model of a SMIB equipped with various FACTS devices. The

transient energy function and the additional damping are also studied.

2.1 Mathematical model of SMIB with and without FACTS device

SMIB without FACTS devices: The SMIB system considered in this study is shown in figure 1. It is used in references [11-12]. It consists of a generator connected to an infinite bus through a step-up transformer and a double circuit transmission line. The system equivalent circuit diagram is given in figure 1.b.

Expressions of reactance's X_1 , X_2 and X are:

$$\begin{aligned} X_1 &= x'_d + x_t \\ X_2 &= \frac{X_{L1} X_{L2}}{X_{L1} + X_{L2}} \\ X &= X_1 + X_2 \end{aligned} \quad (1)$$

The transmission line parameters are:

$$X_{L1} = 0.93 \text{ pu}, X_{L2} = 0.5 \text{ pu}.$$

The generator parameters in per unit are:

$$x'_d = 0.3 \quad H = 3.5 \text{ MW.s / MVA} \quad D = 0 \quad x_t = 0.15$$

The initial system operating conditions are:

$$P = 0.9, Q = 0.436, V_t = 1.0 \angle 28.34^\circ, V_b = 0.90081 \angle 0^\circ$$

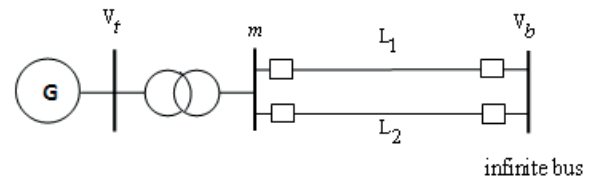


Figure 1. (a) SMIB diagram

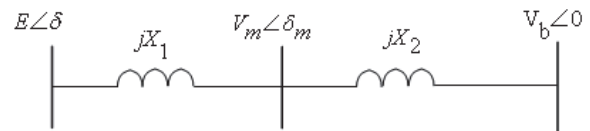


Figure 1. (b) SMIB equivalent circuit

The machine classical electromechanical model is represented by the following differential equations [13]:

$$\begin{aligned} \frac{d\delta}{dt} &= \omega \\ M \frac{d\omega}{dt} &= P_m - P_e - D\omega \end{aligned} \quad (2)$$

For the system without any FACTS devices, the output electrical power P_e is given by:

$$P_e = \frac{EV_b}{X} \sin \delta \quad (3)$$

The generator output power P_e is the main factor that controls the dynamics of the system [14]. In this work, the damping is improved by modulating P_e through two FACTS devices: SSSC and STATCOM.

SMIB with SSSC: The SSSC is a solid-state voltage source inverter connected in series to power transmission



lines [15]. It can vary the effective impedance of a transmission line so as it can influence the power flow by injecting a controllable voltage V_s as shown in figure 2. The injected voltage is in quadrature with the line current and emulates an inductive or a capacitive reactance. For the SSSC the only controllable parameter is the magnitude V_s , note that V_s changes only the current magnitude not its angle.

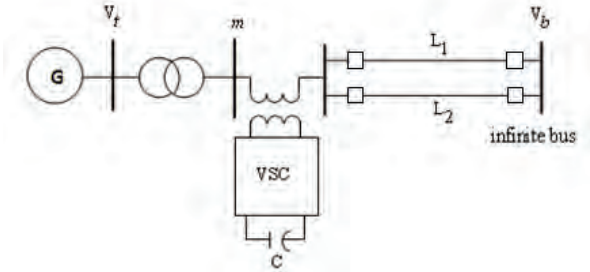


Figure 2. (a) Schematic diagram of SMIB with SSSC

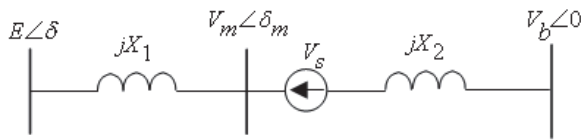


Figure 2. (b) Equivalent circuit of a SMIB with SSSC

In case with SSSC, the current is given by:

$$\bar{I}_s = \frac{\bar{E} - \bar{V}_s - \bar{V}_b}{jX} \quad (4)$$

$\bar{V}_s$ is kept in quadrature with the line current, it is given as follows:

$$\bar{V}_s = V_s e^{j\alpha_s} \quad (5)$$

When V_s lags the line current by 90° the SSSC behaves like a capacitive reactance. In the opposite case, it behaves like an inductive reactance. The phasor diagram of the system with SSSC is shown in figure 3.

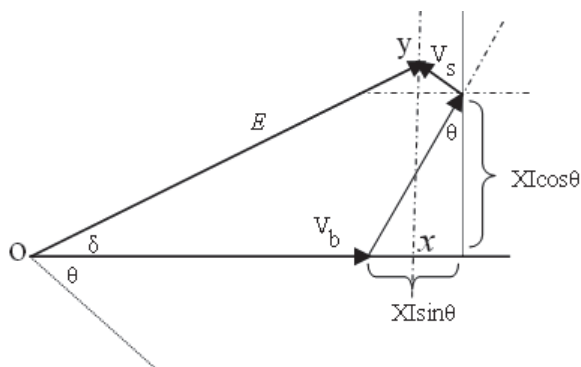


Figure 3. Phasor diagram of a SMIB with SSSC

From figure 3, the terms $XI \cos \theta$ and $XI \sin \theta$ are given by:

$$XI \cos \theta = E \sin \delta - V_s \sin\left(\frac{\pi}{2} - \theta\right) \quad (6)$$

$$XI \sin \theta = E \cos \delta - V_b + V_s \cos\left(\frac{\pi}{2} - \theta\right)$$

After manipulation of equation (6), we get

$$(XI + V_s) \cos \theta = E \sin \delta \quad (7)$$

$$(XI + V_s) \sin \theta = E \cos \delta - V_b$$

Hence θ is expressed as follows:

$$\theta = \tan^{-1} \left(\frac{V_b - E \cos \delta}{E \sin \delta} \right) \quad (8)$$

The electrical power flow P_e is given by:

$$P_e = V_b I \cos \theta \quad (9)$$

From figure 3, we have that

$$xy = XI \cos \theta - V_s \cos \theta = E \sin \delta \quad (10)$$

Thus, the term $I \cos \theta$ is expressed by:

$$I \cos \theta = \frac{E \sin \delta + V_s \cos \theta}{X} \quad (11)$$

P_e is expressed so as follows:

$$P_e = \frac{EV_b}{X} \sin \delta + \frac{VV_s}{X} \cos \theta \quad (12)$$

From equation (7), the term $XI + V_s$ is given by:

$$XI + V_s = \sqrt{E^2 + V_b^2 - 2EV_b \cos \delta} \quad (13)$$

From equation (7) and (13) the term $\cos \theta$ given by:

$$\cos \theta = \frac{E \sin \delta}{\sqrt{E^2 + V_b^2 - 2EV_b \cos \delta}} \quad (14)$$

Equation (12) combined with (14), P_e is rewritten as:

$$P_e = \left(\frac{EV_b}{X} \sin \delta \right) (1 + \xi) \quad (15)$$

$$\text{Where } \xi = \frac{V_s}{\sqrt{E^2 + V_b^2 - 2EV_b \cos \delta}}$$

In the case of SSSC, P_e is modulated as follows: when the generator speed deviation is positive P_e is increased and is decreased when the speed deviation is negative. However, P_e is mainly controlled by the series injected voltage V_s . So, it is positive (negative) when the SSSC operates in capacitive (inductive) mode. Rotor speed deviation ω is considered as control signal; thus V_s can be related dynamically to ω and must satisfy the above criterion [16]:

$$V_s = k_1 \omega; \quad -V_{s\max} \leq V_s \leq V_{s\max} \quad (16)$$

k_1 is a positive gain

SMIB with STATCOM: a STATCOM is a voltage-source converter (VSC) capable of generating or absorbing independently controllable reactive power at its terminal. It is represented by a shunt reactive current I_s source as shown in figure 4.a [17]. STATCOM is placed in bus m. its reactive current is always kept in



quadrature with its terminal voltage and can be written for capacitive mode as:

$$\bar{I}_s = I_s e^{j(\delta_m - 90^\circ)} \quad (17)$$

For an inductive mode of operation, I_s in (17) is replaced by $-I_s$.

From figure 4.b the complex current $\bar{I}$ and the complex voltage in the intermediate bus m are given by:

$$\bar{I} = \frac{\bar{E} - \bar{V}_m}{jX_1} \quad (18)$$

$$\bar{V}_m = \bar{V}_b + jX_2(\bar{I} - \bar{I}_s)$$

Replacing in (18), $\bar{I}$ with its expression, $\bar{V}_m$ can be written as follows:

$$\bar{V}_m = \frac{X_1 V_b}{X} + \frac{X_2 E e^{j\delta}}{X} - \frac{X_2 X_1 I_s e^{j(\delta_m \pm 90^\circ)}}{X} \quad (19)$$

From (19), voltage magnitude and angle of bus m are given by:

$$V_m = \frac{EX_2 \cos(\delta - \delta_m) + V_b X_1 \cos \delta_m + X_1 X_2 I_s}{X}$$

$$\delta_m = \tan^{-1} \left(\frac{EX_2 \sin \delta}{VX_1 + EX_2 \cos \delta} \right) \quad (20)$$

Replacing in (18) $\bar{V}_m$ with its expression, the complex current $\bar{I}$ can be written as:

$$\bar{I} = \frac{E e^{j\delta} - V_b}{jX} + \frac{X_2}{X} I_s e^{j(\delta_m - 90^\circ)} \quad (21)$$

P_e is expressed as: $P_e = \text{real}(\bar{E} \bar{I}^*)$

Replacing $\bar{I}$ by its expression in (21), P_e is generalized as:

$$P_e = \frac{EV_b}{X} \sin \delta + \frac{EX_2}{X} \sin(\delta - \delta_m) I_s \quad (22)$$

In the case of STATCOM, P_e is modulated by the injected or absorbed shunt current I_s which is related dynamically to rotor speed deviation ω . Thus, I_s satisfies the above criterion:

$$I_s = k_2 \omega; \quad -I_{s\max} \leq I_s \leq I_{s\max} \quad (23)$$

k_2 is a positive gain

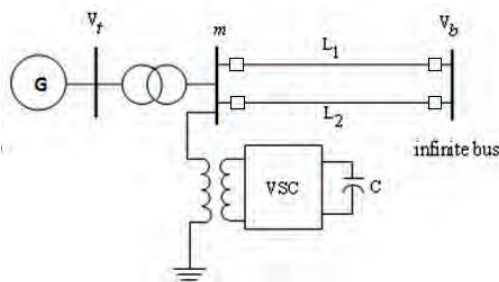


Figure 4. (a) Schematic diagram of SMIB with STATCOM

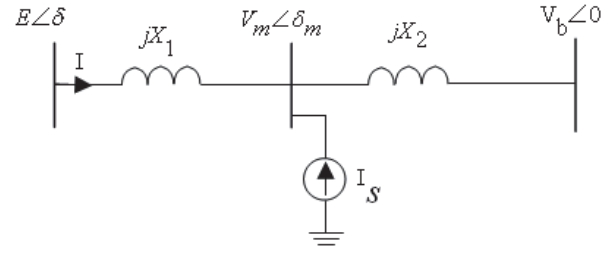


Figure 4. (b) Equivalent circuit of a SMIB with STATCOM

2.2. Transient energy function and additional damping:

The machine classical electromechanical model at the post fault stable equilibrium point is represented by the following differential equations [13]:

$$\frac{d\delta_s}{dt} = \omega_s \quad (24)$$

$$M \frac{d\omega_s}{dt} = P_m - P_e - D\omega_s$$

At the post-fault stable equilibrium point, the mechanical power is equal to electrical power so, δ_s is given as follows:

$$\delta_s = \sin^{-1} \left(\frac{P_m}{\frac{EV_b}{X}} \right) \quad (25)$$

By the transformation $x = \delta - \delta_s$, (2) combined with (24) and (25), we obtain:

$$M \frac{d^2 x}{dt^2} + M \frac{d^2 \delta_s}{dt^2} + D \frac{dx}{dt} + D \frac{d\delta_s}{dt} = \frac{EV_b}{X} (\sin \delta_s - \sin(x + \delta_s)) \quad (26)$$

Specifying the state variables x_1 and x_2 as follows:

$$x = x_1 \quad \dot{x} = \dot{\delta} = \omega = x_2$$

Equation (26) can be rewritten as follows:

$$\frac{dx_1}{dt} = x_2 \quad (27)$$

$$\frac{dx_2}{dt} = \frac{EV_b}{MX} (\sin \delta_s - \sin(x_1 + \delta_s))$$

From (27), we obtain:

$$x_2 \frac{dx_2}{dt} = \frac{EV_b}{MX} (\sin \delta_s - \sin(x_1 + \delta_s)) \frac{dx_1}{dt} \quad (28)$$

By Integrating (28), we obtain:

$$\frac{1}{2} x_2^2 = -\frac{1}{M} \int_0^{x_1} \frac{EV_b}{X} (-\sin \delta_s + \sin(u + \delta_s)) \frac{du}{dt} \quad (29)$$

Defining the transient energy function $V(x_1, x_2)$ as follows:

$$V(x_1, x_2) = \frac{1}{2} x_2^2 + \frac{1}{M} \int_0^{x_1} \frac{EV_b}{X} (-\sin \delta_s + \sin(u + \delta_s)) \frac{du}{dt} \quad (30)$$

$V(x_1, x_2)$ can be rewritten as [5]:

$$V(x_1, x_2) = \frac{1}{2} M \omega^2 + [-P_m(\delta - \delta_s) - \frac{EV_b}{X} \cos(\delta - \delta_s)] \quad (31)$$



Physically, $\frac{1}{2} M \omega^2$ is called the kinetic energy V_{KE} , it is function of rotor speed deviation ω .

$-P_m(\delta - \delta_s) - \frac{EV_b}{X} \cos(\delta - \delta_s)$ is called the potential

energy V_{PE} . It is function of rotor angle δ .

The transient energy given by equation (31) can be written as follows [14]:

$$V(\delta, \omega) = V_{KE} + V_{PE}$$

The time derivative of energy function of 24 can be written as follows [14].

$$\begin{aligned} \dot{V} &= \frac{dV}{dt} = \frac{dV_{KE}}{dt} + \frac{dV_{PE}}{dt} \\ &= M \omega \frac{d\omega}{dt} - P_m \frac{d\delta}{dt} + \frac{EV_b}{X} \sin \delta \frac{d\delta}{dt} \end{aligned} \quad (32)$$

Replacing in equation (32) ω and δ with those in equation 2, the transient energy function is rewritten as:

$$\dot{V} = -(D\omega^2 + P_e \omega - \frac{EV_b}{X} \omega \sin \delta) \quad (33)$$

For the case of SMIB without FACTS devices, the transient energy function can be written as [6]:

$$\dot{V} = -D\omega^2 \quad (34)$$

For the case with SSSC, the transient energy function can be written as [13]:

$$\dot{V} = -(D + k_1 \xi \frac{EV_b}{X} \sin \delta) \omega^2 \quad (35)$$

For the case with STATCOM, the transient energy function can be written as [13]:

$$\dot{V} = -(D + k_2 \frac{EX_2}{X} \sin(\delta - \delta_m)) \omega^2 \quad (36)$$

D is the natural damping coefficient, for the case with FACTS devices; a new term is added and can be considered as an additional damping.

So we can define D_{SSSC} and D_{STAT} as additional damping provided by SSSC and STATCOM respectively as follows:

$$D_{SSSC} = k_1 \xi \frac{EV_b}{X} \sin \delta \quad (37)$$

$$D_{STAT} = k_2 \frac{EX_2}{X} \sin(\delta - \delta_m)$$

The equivalent circuit of the system without FACTS in the prefaulted period is the same as in figure 1.b. The reduced equivalent circuits during faulted and post-faulted period are shown in figure 5.

In the pre-fault period, the electrical power is given by equation (3), but during faulted and post-faulted period the electrical powers are expressed respectively as follows:

$$P_{efault} = 0$$

$$P_{epost-fault} = \frac{EV_b}{X_1 + X_{L1}} \sin \delta \quad (37)$$

For the case with SSSC, The equivalent circuit of the system in the prefaulted period is the same as in figure 2.b. the reduced equivalent circuits during faulted and post-faulted period are shown in figure 6.

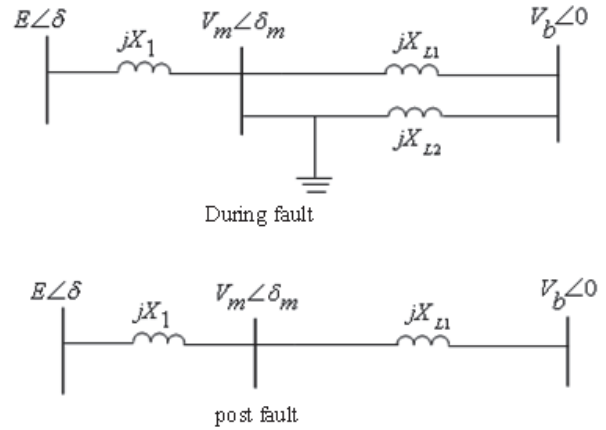


Figure 5. Reduced equivalent circuit of a SMIB without FACTS

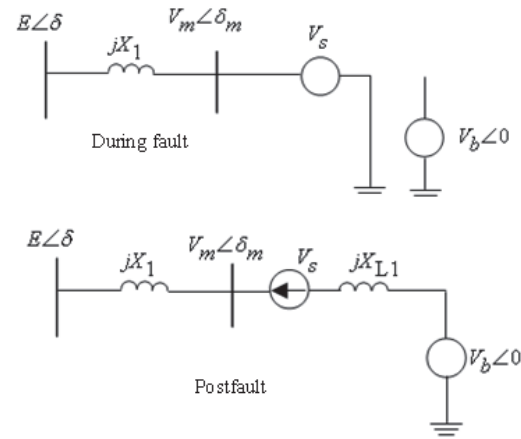


Figure 6. Reduced equivalent circuit of a SMIB with SSSC

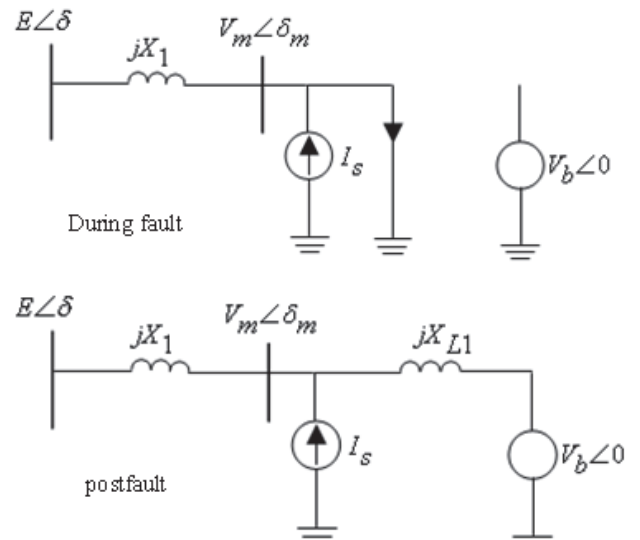


Figure 7. Reduced equivalent circuit of a SMIB with STATCOM

The electrical power in the pre-faulted period is given by equation (15), but during faulted and post faulted period the electrical powers are expressed respectively as follows:



$$P_{efault} = \frac{EV_s}{X_1 + X_{L1}} \cos(\delta - \theta)$$

$$P_{epostflt} = \left(\frac{EV_b}{X_1 + X_{L1}} \sin \delta \right) (1 + \xi) \quad (38)$$

For the case with STATCOM, The equivalent circuit of the system in the prefaulted period is the same as in figure 4.b.

The reduced equivalent circuits during faulted and post-faulted period are shown in figure 7.

The electrical power in the pre-faulted period is given by equation (22), but during faulted and post faulted period the electrical powers are expressed respectively as follows:

$$P_{epfault} = EI_s \cos(\delta - \delta_m \pm \frac{\pi}{2})$$

$$P_{epostflt} = \frac{EV_b}{X} \sin \delta + \frac{EX_{L1}}{X_1 + X_{L1}} \sin(\delta - \delta_m) I_s \quad (39)$$

Where δ_m is given by:

$$\delta_m = \tan^{-1} \left(\frac{EX_{L1} \sin \delta}{VX_1 + EX_{L1} \cos \delta} \right)$$

3. Approach of injection model

3.1 SSSC with injection model

Consider the configuration of SSSC inserted between bus m and bus n of power system as shown in Figure 8.a. Two consuming buses (loads): bus i and bus j are added, their fictitious active and reactive power injection are calculated by equations 40 and 41 respectively [10].

$$P_i = rbV_i^2 \sin \gamma \quad (40)$$

$$Q_i = rbV_i^2 \cos \gamma$$

$$P_j = -rbV_jV_j \sin(\theta_i - \theta_j + \gamma) \quad (41)$$

$$Q_j = -rbV_iV_j \cos(\theta_i - \theta_j + \gamma)$$

Where $0 \leq r \leq r_{\max}$, $0 \leq \gamma \leq 2\pi$ and $b = \frac{1}{X_s}$

r and γ are magnitude and angle of the injected voltage V_s in the case of injection model.

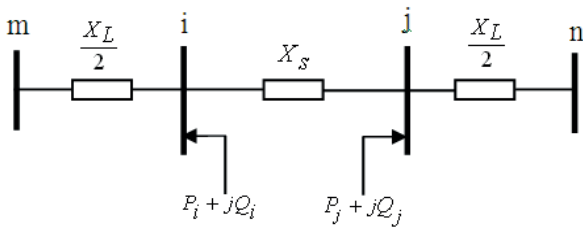


Figure 8.a. Diagram of SSSC with injected model.

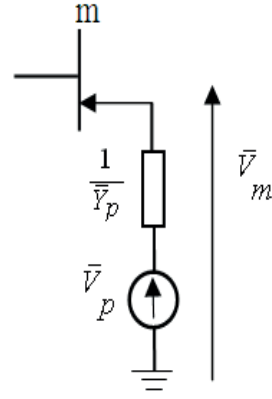


Figure 8.b. Diagram of STATCOM with injected model.

3.2 STATCOM with injection model

Consider the configuration of STATCOM inserted at bus m of power system as shown in Figure 8.b.

The fictitious active and reactive power injection are calculated by equation 42.

$$P_m = V_m^2 Y_p \cos \theta_p - Y_p V_p V_m \cos(\alpha_m - \alpha_p - \theta_p) \quad (42)$$

$$Q_m = -V_m^2 Y_p \sin \theta_p - Y_p V_p V_m \sin(\alpha_m - \alpha_p - \theta_p)$$

3.3 Electrical network modelling

The swing model of a generator in electrical network is given by [18]:

$$M_i \ddot{\delta}_i + D \dot{\delta}_i = P_{mi} - P_{ei} \quad (43)$$

The electrical power of the i-th generator is given:

$$P_{ei} = \sum_{j=1}^n E_i E_j Y_{ij} \cos(\delta_i - \delta_j - \theta_{ij}) \quad (44)$$

4. Simulations and results

4.1 The single machine infinite bus

The SMIB is described in section 2, its diagram and equivalent circuit are shown in figure 1.

Note that we have implemented the first approach for the SMIB. two cases of studies are presented.

Case 1: In this case a 10% step disturbance in the mechanical power P_m of the machine is applied. Rotor angle and speed deviation responses are given in figure 10 and 11 respectively.

It's clear that without FACTS devices there are maintained oscillations with poor damping of rotor angle and speed deviation.

With STATCOM, the rotor angle and speed deviation response show that oscillations take more than 5s to be damped out completely.

With SSSC, The rotor angle and speed deviation take only about 2s to reach their steady state values and oscillations are so completely damped out. This decrease in oscillations is related to the transient energy response as shown in figure 12.



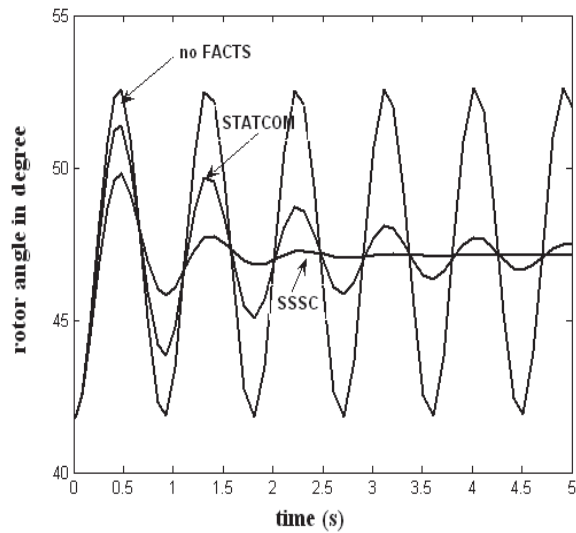


Figure 10. Rotor angle response

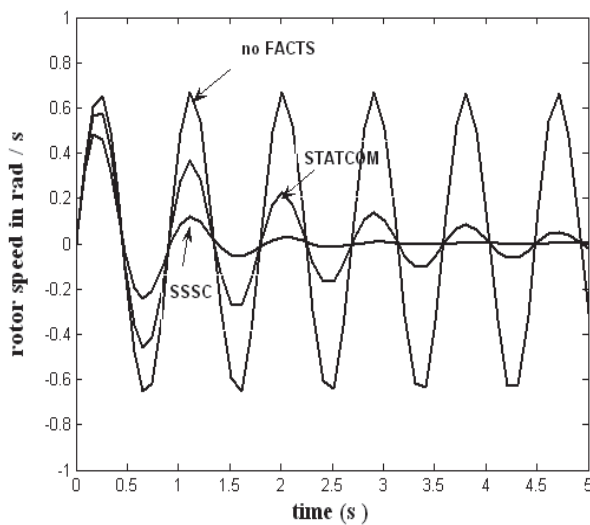


Figure 11. Rotor speed deviation response

It is clear, that without any FACTS devices, the transient energy oscillations are kept constant during disturbance. However, when we added a STATCOM, oscillations decrease and the transient energy takes about 3s to be reach it's pre-disturbance value (which is equal to zero), when we added a SSSC to the SMIB, the transient energy decreases rapidly and takes only 0.5s to be zero and oscillations are rapidly damped this indicates the addition of some damping to the system with FACTS devices as shown in figure 13, it's clear that the SSSC provides more additional damping than the STATCOM that's why energy function reaches rapidly the zero value with SSSC.

Figure 14 shows the SSSC injected voltage and the STATCOM injected current responses to the considered disturbance, here $I_{s\max} = 0.05$ and $V_{s\max} = 0.1$.

Values of $I_{s\max}$ and $V_{s\max}$ are limited by flow of reactive power of the STATCOM and flow of active power of the SSSC and line transmission parameters, these values are generally small and lower than 0.25 pu.

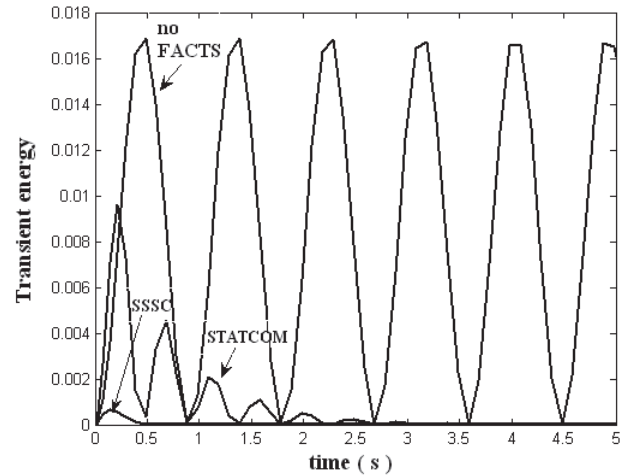


Figure 12. Transient energy response

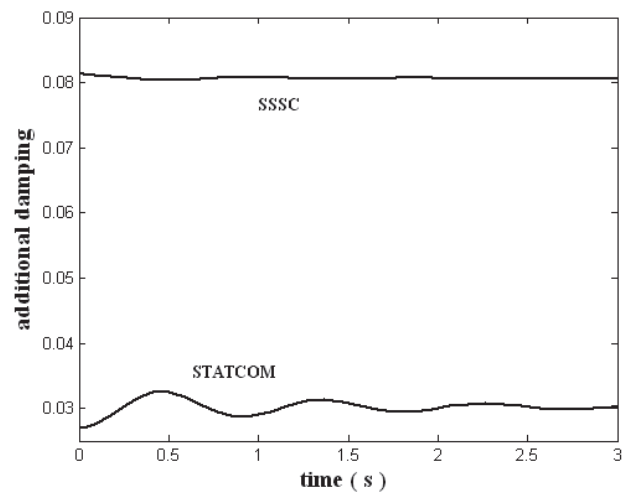


Figure 13. Additional damping with FACTS devices

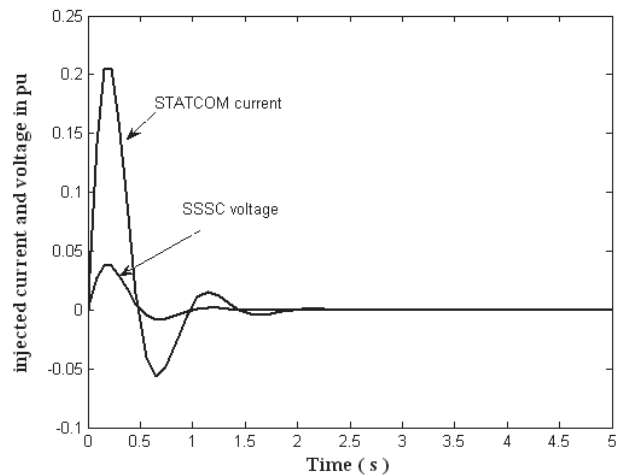


Figure 14. SSSC injected voltage and STATCOM injected current

Case2:

A three phase fault occurs at bus m at $t=1s$ and is cleared by disconnecting line 2 at end. Rotor angle deviation response to this disturbance is shown in figure 15.

From figure we see that the system is stable with a critical clearing time $t_c = 0.086s$, and is unstable with $t_c=0.087s$. The critical clearing time is therefore $0.0865+0.0005s$. for a critical clearing time $t_c= 0.088s$, we have plotted the same curves as in case1 for the SMIB with and without FACTS devices.



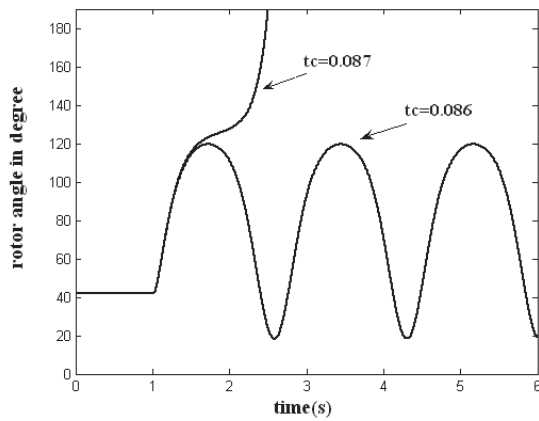


Figure 15. Rotor angle response for different clearing time

For an initial generator output power of 0.90 pu, the angle δ in the pre-fault and post fault stable equilibrium point is found as 41.7714 and 54.7260°, respectively. Figure 16 and figure 17 show curves of machine rotor angle and machine rotor speed plotted without FACTS devices, with STATCOM and SSSC.

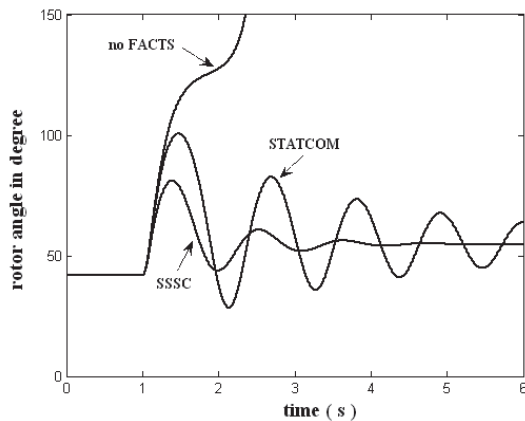


Figure 16. Rotor angle response with and without FACTS

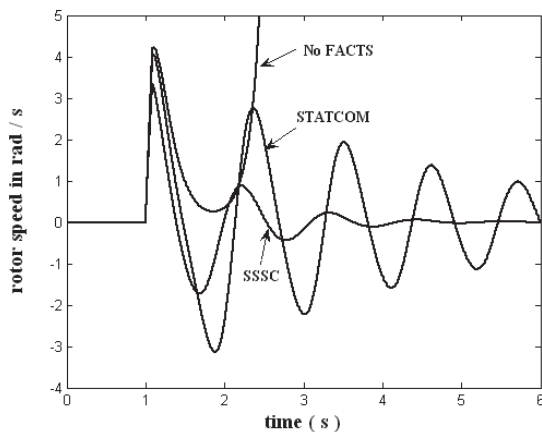


Figure 17. Rotor speed response with and without FACTS

It's clear, at $t_c = 0.088$ s, that the system is unstable in the absence of FACTS, but when we added the STATCOM current, the system remains stability, oscillations are damped out and the damping is significantly improved. This later can further be improved by adding the SSSC voltage. The transient energy function is shown in figure 18, it's clear that without FACTS, the transient energy is constant in the

pre-faulted period, then it increases rapidly and oscillates severely during faulted period and sustained constant at the post fault period. When we added STATCOM, it oscillates during faulted period but decreases rapidly to reach its pre-fault value in about 6s during post-fault period. However when we added SSSC, it oscillates during faulted period but regains rapidly its prefault value in about 2s during post – faulted period.

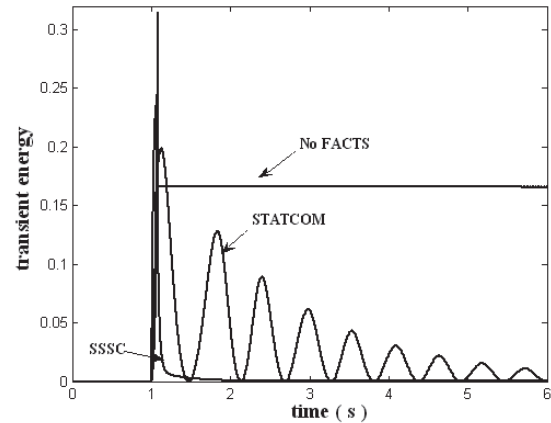


Figure 18. Transient energy response with and without FACTS

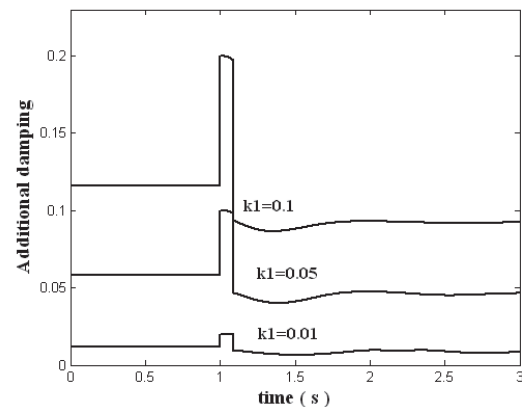


Figure 19. SSSC additional damping

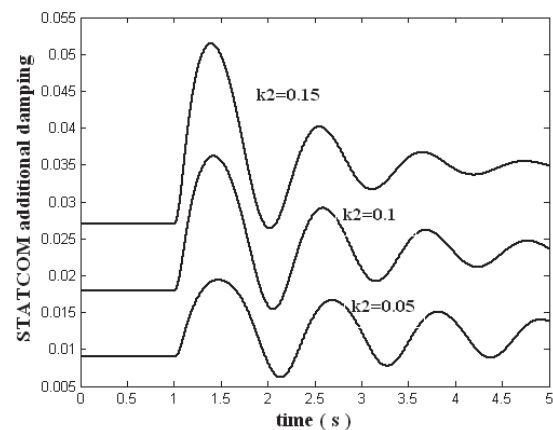


Figure 20. STATCOM additional damping

The additional damping provided by the SSSC and STATCOM for various values of k_1 and k_2 is shown in figure 19 and figure 20 respectively.

It's clear that the SSSC damping D_{SSSC} increases as the value of k_1 increases, so higher value of k_1 (hence



D_{SSSC}) involves a decrease in the system with SSSC transient energy function.

The STATCOM damping D_{STAT} increases as the value of k_2 increases, so higher value of k_2 (hence D_{STAT}) involves a decrease in the system with STATCOM transient energy function.

Figure 21 shows the controls signals: V_s voltage of SSSC and I_s reactive current of STATCOM, here $V_{smax} = 0.1$ and $I_{smax} = 0.1$.

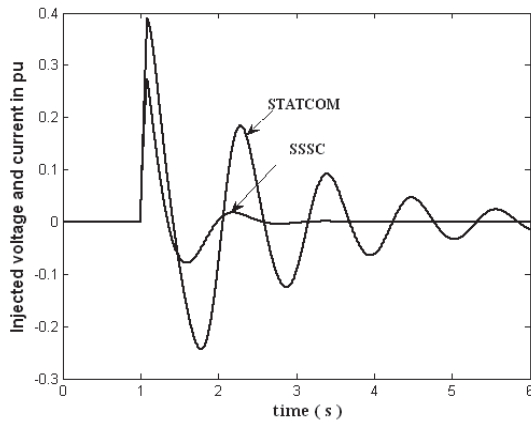


Figure 21. SSSC injected voltage and STATCOM current

4.2 The multimachine power system

The multimachine power system studied in this work is used in reference [18], its diagram is given by figure 22.

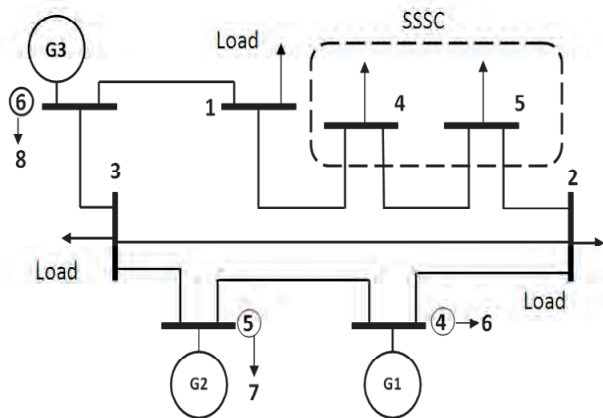


Figure 22. Diagram of the considered system.

A fault appears in bus 2 at $t=0.1s$ and cleared at $t=0.2s$, figure 23 shows curves of machine rotor angles with respect to machine 4 for the uncompensated case. Then, an SSSC and a STATCOM modeled with the approach of injection model are inserted to network. SSSC is placed in the middle of line (1-2), the structure of network is modified as seen in figure 22; two new additional consumed buses 4 and 5 are added. The STATCOM is connected at bus 1, the network structure was not modified.

Here, we have chosen $b = 14.3$; $r = 0.15$; $\gamma = \pi / 6$

and $V_p = 1.02$, $\alpha_p = -\frac{\pi}{6}$, $\theta_p = -\frac{\pi}{2}$, $Y_p = 1 / 0.625$.

Figure 24, figure 25 and figure 26 shows, respectively, curves of rotor angles of machine 1, machine 2 and machine 3 with respect to machine 4 for the cases network with SSSC, network with STATCOM and uncompensated network.

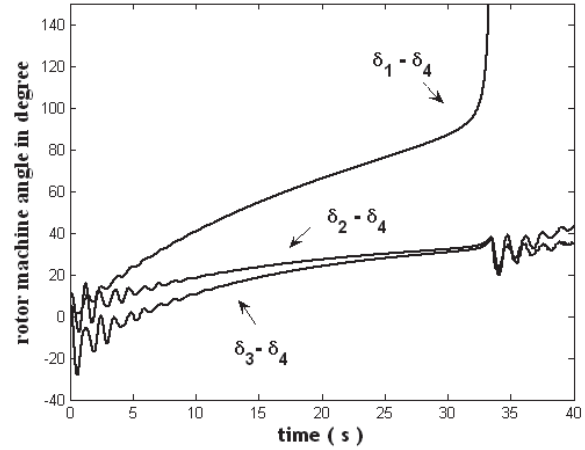


Figure 23. curves of machines with respect to machine 4

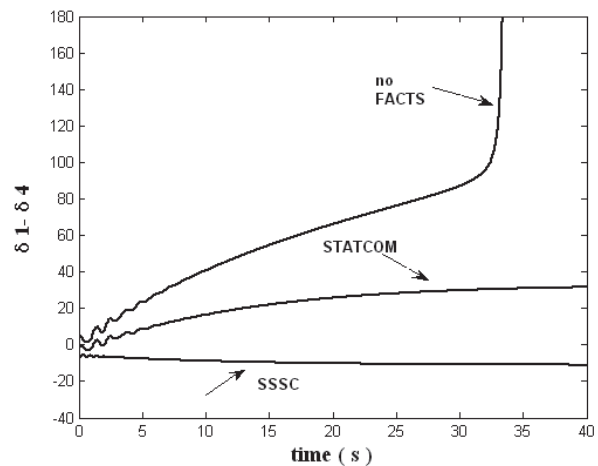


Figure 24. rotor angle of machine 1 with respect to machine 4

From figures, it can be seen, for the uncompensated case, that machine 1 lose synchronism and the system is instable, when we insert SSSC and STATCOM, rotor angle of machine 1 with respect to 4 regains synchronism and the entire system is so stable, oscillations are further damped out as shown in figure 24.

Curves of machine 2 and 3 with respect to machine 4, shows that machines do not lose synchronism but, rotor angle oscillated severely, when we insert SSSC and STATCOM, oscillations are suppressed.

It can be concluded, that SSSC and STATCOM implemented with the two strategies, for the SMIB and a multimachine system are efficient in damping oscillations and remaining stability under various several disturbances.

It's noted either, that the SSSC reacts further than STATCOM and damped out more oscillations in a short time.



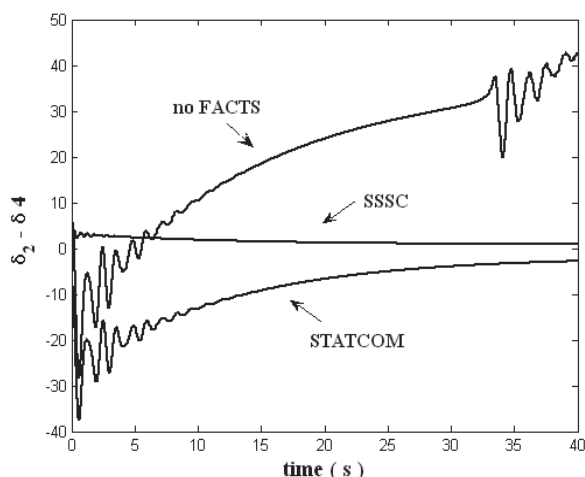


Figure 25. rotor angle of machine 2 with respect to machine 4

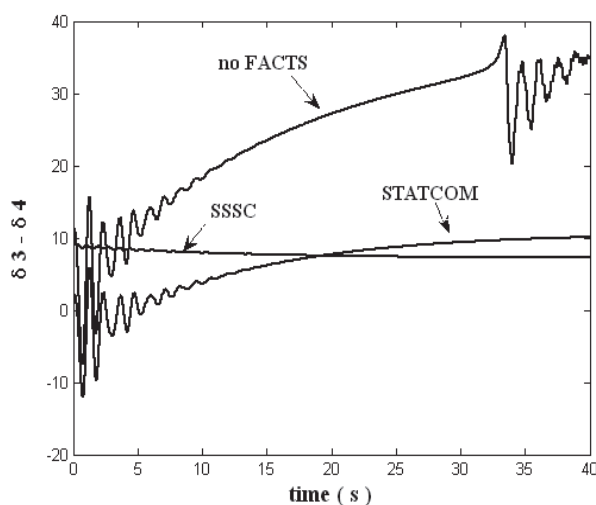


Figure 26. rotor angle of machine 3 with respect to machine 4

3. Conclusion

In this paper, the goal of work is to show the influence of FACTS devices on power system stability. The concept that transient energy is a tool to measure dynamic system damping is exploited in this work. Another strategy based on injection model was implemented to demonstrate these FACTS devices effect on multi machine power system stability. Analytical expressions of transient energy and additional damping provided by two types of FACTS devices are presented. Mathematical model of the injection model was also derived.

The two strategies are runned for both a single and multimachine power system under several perturbations. The SSSC performs a faster oscillations damping improvement then STATCOM hence provided a better additional damping and this accelerates transient energy function to reach zero rapidly and the system to regain its stable post fault equilibrium point.

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Design of Fuzzy Logic Controller for SSSC and TCPS based Automatic Generation Control of Hydrothermal System under Deregulated Environment

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Abstract

This paper presents the modelling and simulation of Fuzzy Logic Controller (FLC) for Static Synchronous Series Compensator (SSSC) based Automatic Generation Control (AGC) under deregulated environment. The combined effect of SSSC and Thyristor Controlled phase Shifter (TCPS) is also carried out. A two area hydrothermal system under deregulated environment has been considered for this purpose. The devices are modeled and attempt has been made to incorporate these devices in the two area system thus improving the dynamic response of the system. The effect of these parameters on the system is demonstrated with the help of computer simulations. A systematic method has also been demonstrated for the design of FLC. Computer simulations reveal that due to the presence of Fuzzy logic controller, the dynamic performance of the system in terms of settling time, overshoot is greatly improved than that of without FLC.

Keywords: Fuzzy Logic Controller, Static Synchronous Series Compensator, TCPS, Deregulation.

1. Introduction

Successful operation of a power system is the process of properly maintaining several sets of balances. Two of these balances are between load-generation and scheduled and actual tie line flows. These two balances are predominant factors to keep frequency constant. Constant frequency is identified as the primary index of healthy operation of system and the quality of supplied power to consumer as well. Both of these balances are maintained by adjusting generation keeping load demand in view. If frequency is low, generation is increased and if the actual outflow is greater than the scheduled outflow, generation is decreased. Since system

conditions are always changing as load constantly varies during different hours of a day, precise manual control of these balances would be impossible. Automatic Generation Control (AGC) was developed to both maintain a (nearly) constant frequency and to regulate tie line flows [1-3].

Under open market system (deregulation) the power system structure changed in such a way that would allow the evolving of more specialized industries for generation (Genco), transmission (Transco) and distribution (Disco). A detailed study on the control of generation in deregulated power systems is given in [4]. The concept of independent system operator (ISO) as an unbiased coordinator to balance reliability with economics has also emerged [5-6]. The assessment of Automatic Generation control in a deregulated environment is given in detail in [7] and also provides a detailed review over this issue and explains how an AGC system could be simulated after deregulation.

On the other hand, the concept of utilizing power electronic devices for power system control has been widely accepted in the form of Flexible AC Transmission Systems (FACTS) which provide more flexibility in power system operation and control [8]. A Static Synchronous Series Capacitor is expected to be an effective apparatus for the tie-line power flow control of an interconnected power system in the analysis of an interconnected power system. The proposed control strategy will be a new ancillary service for the stabilization of frequency oscillations of an interconnected power system.

A few investigations have been carried out using Fuzzy Logic Controllers (FLC) for AGC of thermal systems [9]. Surprisingly, till date, no attempt has been made to examine the effect of FLC in an interconnected hydro-thermal system with Thyristor controlled phase shifter under deregulated environment.

The remainder of the paper is organized as follows: Section (2) focuses on the dynamic mathematical model considered in this work. Section (3) emphasizes on the



modelling of SSSC and TCPS. The design of FLC is presented in Section (4). Section (5) presents the

detailed analysis of results and conclusions are drawn in Section (6)

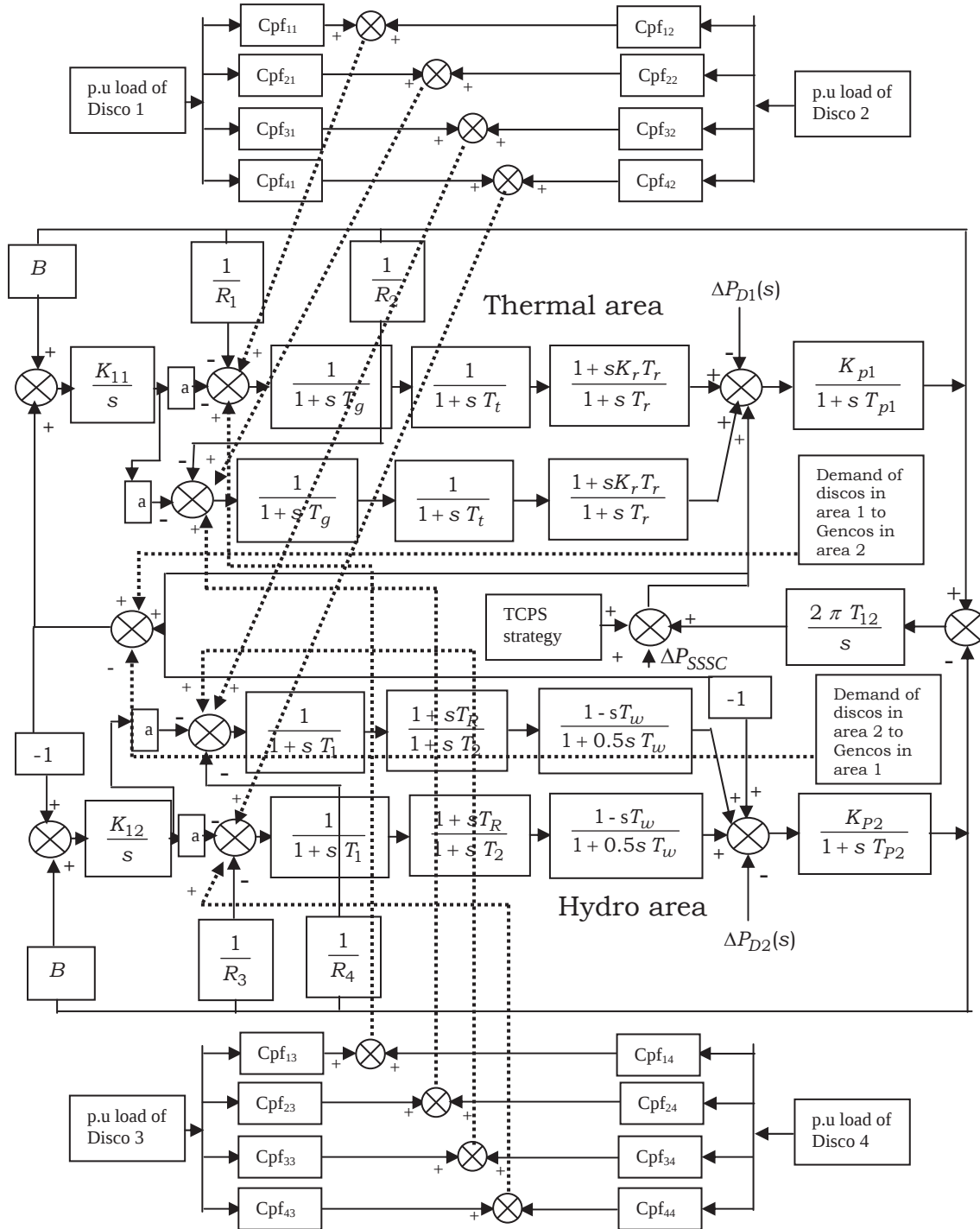


Figure 1. Two Area hydrothermal system with TCPS and SSSC

2. Dynamic Mathematical Model

The Automatic Generation Control system investigated is composed of an interconnection of two areas under deregulated scenario. Area 1 comprises of a reheat system and area 2 comprises of hydro system. Figure. 1 is the block diagram of two-area hydrothermal system under deregulated scenario where ACE of each area is fed to the corresponding controller. The accurate control

signal is generated for every incoming ACE at that particular load change. A performance index given by

$$J = \int_0^t (\alpha \cdot \Delta f_1^2 + \beta \cdot \Delta f_2^2 + \Delta P_{tie12}^2) dt$$

to compare the performance of Fuzzy logic controller and integral controller. Here the values of α and β are considered to be equal to 0.065.



3. Modelling of SSSC applied to AGC

A SSSC employs self-commutated voltage-source switching converters to synthesize a three-phase voltage in quadrature with the line current, emulates an inductive or a capacitive reactance so as to influence the power flow in the transmission lines. The compensation level can be controlled dynamically by changing the magnitude and polarity of injected voltage, V_s and the device can be operated both in capacitive and inductive mode. The schematic of an SSSC, located in series with the tie-line between the interconnected areas, can be applied to stabilize the area frequency oscillations by high speed control of the tie-line power through interconnection as shown in Figure.

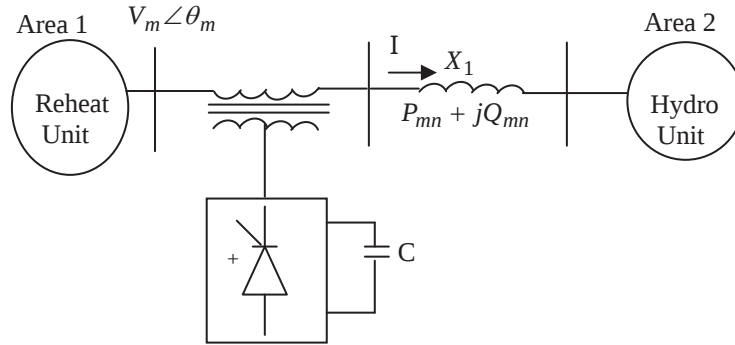


Figure 2: Schematic of SSSC applied to AGC

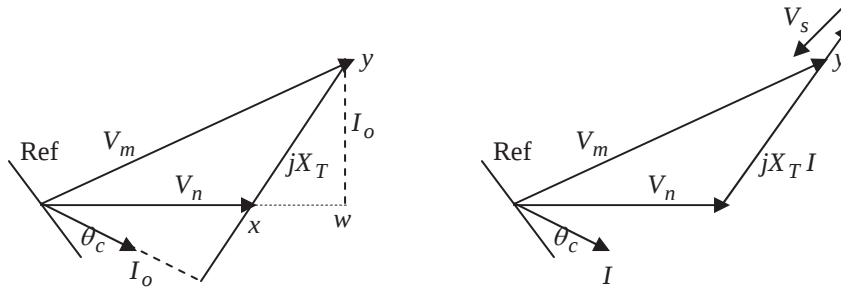


Figure 3: Phasor diagram at $V_s=0$ and V_s lagging I by 90°

Where $X_T = X_L + X_S$. The phase angle of the current can be expressed as

$$\theta_c = \tan^{-1} \left[\frac{V_n \cos \theta_n - V_m \cos \theta_m}{V_m \sin \theta_m - V_n \sin \theta_n} \right] \quad (2)$$

But Eqn (1) can be expressed in a generalized form as

$$I = \frac{V_m - V_s - V_n}{jX_T} = \left[\frac{V_m - V_n}{jX_T} \right] + \left[\frac{-V_s}{jX_T} \right] = I_o + \Delta I \quad (3)$$

The term ΔI is an additional current term due to SSSC voltage V_s . The power flow from bus m to bus n can be written as

$$S_{mn} = V_m I^* = S_{mno} + \Delta S_{mn} \text{ which implies} \\ P_{mn} + jQ_{mn} = (P_{mno} + \Delta P_{mn}) + j(Q_{mno} + \Delta Q_{mn}) \quad (4)$$

2. The equivalent circuit of the system shown in Figure 2 can also be represented by a series connected voltage source V_s along with a transformer leakage reactance X_s . The SSSC controllable parameter is V_s , which in fact represents the magnitude of injected voltage. Figure 3 represents the phasor diagram of the system taking into account the operating conditions of SSSC.

Based on the above figure when $V_s = 0$, the current I_o of the system can be written as

$$I_o = \frac{V_m - V_n}{jX_T} \quad (1)$$

Where P_{mno} and Q_{mno} are the real and reactive power flow respectively when $V_s = 0$. The change in real power flow caused by SSSC voltage is given by

$$\Delta P_{mn} = \frac{V_m V_s}{X_T} \sin(\theta_m - \alpha) \quad (5)$$

When V_s lags the current by 90° , ΔP_{mn} can be written as

$$\Delta P_{mn} = \frac{V_m V_s}{X_T} \cos(\theta_m - \theta_c) \quad (6)$$

From Eqn (2) the term $\cos(\theta_m - \theta_c)$ can be written as

$$\cos(\theta_m - \theta_c) = \frac{V_n}{V_m} \cos(\theta_n - \theta_c) \quad (7)$$

Referring to Fig 3 it can be written as

$$\cos(\theta_n - \theta_c) = \frac{yw}{xy} \quad (8)$$

$$\text{And it can be seen as } yw = V_m \sin \theta_{mn} \quad (9)$$



Also $xy = \sqrt{V_m^2 + V_n^2 - 2V_m V_n \cos \theta_{mn}}$ and

$$\theta_{mn} = \theta_m - \theta_n \quad (10)$$

Using these relationships Eqn (6) can be modified as follows

$$\Delta P_{mn} = \frac{V_m V_n}{X_T} \sin \theta_{mn} \times \frac{V_s}{\sqrt{V_m^2 + V_n^2 - 2V_m V_n \cos \theta_{mn}}} \quad (11)$$

From Eqn (4) it can be written as

$$P_{mn} = P_{mno} + \Delta P_{mn}$$

Which implies

$$P_{mn} = \frac{V_m V_n}{X_T} \sin \theta_{mn} + \left(\frac{V_m V_n}{X_T} \sin \theta_{mn} \times \frac{V_s}{\sqrt{V_m^2 + V_n^2 - 2V_m V_n \cos \theta_{mn}}} \right) \quad (12)$$

Linearizing Eqn (12) about an operating point it can be written as

$$\Delta P_{mn} = \frac{V_m V_n}{X_T} \cos(\theta_m - \theta_n)(\Delta \theta_m - \Delta \theta_n) + \left(\frac{V_m V_n}{X_T} \sin \theta_{mn} \times \frac{\Delta V_s}{\sqrt{V_m^2 + V_n^2 - 2V_m V_n \cos \theta_{mn}}} \right) \quad (13)$$

$$\Delta P_{mn} = \Delta P_{tie} + \Delta P_{SSSC} \text{ which implies}$$

$$\Delta P_{SSSC} = \left(\frac{V_m V_n}{X_T} \sin \theta_{mn} \times \frac{\Delta V_s}{\sqrt{V_m^2 + V_n^2 - 2V_m V_n \cos \theta_{mn}}} \right) \quad (14)$$

Based on Eqn (14) it can be observed that by varying the SSSC voltage ΔV_s , the power output of SSSC can be controlled which will in turn control the frequency and tie line deviations. The structure of SSSC to be incorporated in the two area system in order to reduce the frequency deviations is provided in Figure 4 shown below. The frequency deviation of area 1 can be seen as input to the SSSC device.

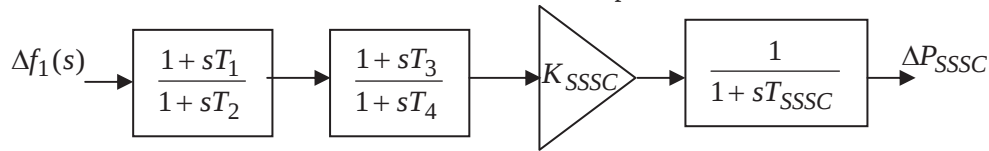


Figure 4: Structure of SSSC

Modelling of TCPS

Figure 5 shows the schematic of the two-area interconnected hydrothermal system considering a TCPS in series with the tie-line. TCPS is placed

near area 1. Area 1 is the thermal area comprising of three reheat units and area 2 is the hydro area consisting of three hydro units. With TCPS, the incremental tie-line power flow from area 1 to area 2 under open market system can be expressed as (16)

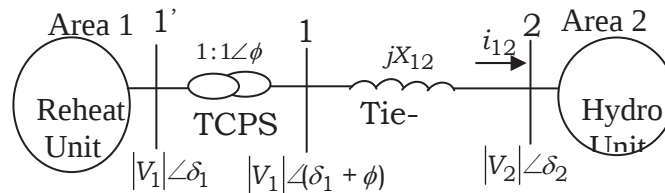


Figure.5. Schematic of TCPS in series with Tie line

The phase shifter angle $\Delta \phi(s)$ can be written as

$$\Delta \phi(s) = \frac{K_\phi}{1 + sT_{ps}} \Delta Error_1(s) \quad (15)$$

Where K_ϕ and T_{ps} are the gain and time constants of the TCPS and $\Delta Error_1(s)$ is the control signal which controls the phase angle of the phase shifter. Thus, it can be written as

$$\Delta P_{tie12}(s) = \frac{2\pi T_{12}'}{s} [\Delta F_1(s) - \Delta F_2(s)] + T_{12}' \frac{K_\phi}{1 + sT_{ps}} \Delta Error_1(s) \quad (16)$$

$\Delta Error_1$ can be any signal such as the thermal area frequency deviation Δf_1 or hydro area frequency deviation Δf_2

4. Fuzzy Logic Controller

The Fuzzy Logic serves as a basis for constructing a set of fuzzy if-then rules with appropriate membership functions to generate the stipulated input-output pairs. In this work a PI-like Fuzzy Knowledge Based Controller has been designed. The basic structure of the conventional PI controller is given by

$$u = K_p e + K_I \int e dt \quad \dots (17)$$

where K_p and K_I are the proportional and integral gains respectively and e is the error signal(i.e.



e =process set point–process output variable). Taking the derivative with respect to time, the above equation 1 is transformed into the following equivalent expression

$$\dot{u} = K_p \dot{e} + K_I e \quad \dots(18)$$

For the Automatic generation control problem the inputs to the Fuzzy controller for i th area at a

particular instant ‘ t ’ are $ACE_i(t)$ and $\Delta ACE_i(t)$, where $ACE_i(t) = \Delta P_{tiei} + B_i \Delta f_i$ and $\Delta ACE_i(t) = ACE_i(t) - ACE_i(t-1)$ and output of the fuzzy controller is Δu . The structure of the FLC resembles that of a knowledgebased controller except that the FLC utilizes the principles of fuzzy set theory in its data representation and its logic. Generally the fuzzy inference system can be shown as shown in Figure 6

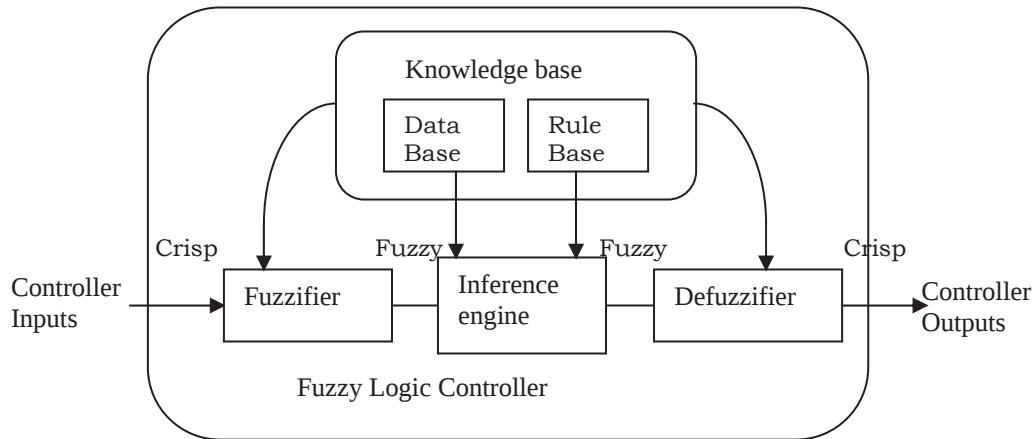


Figure 6: Schematic diagram of the FLC building blocks

5. Results and Discussions

The proposed Fuzzy Logic controller is applied to a two area hydrothermal thermal system under deregulated scenario. The simulation has been conducted with Fuzzy Logic Toolbox and Simulink in MATLAB 7.1 using Mamdani type fuzzy inference system. Three generators in each area have been considered for the study. Each Genco participates in AGC as defined by following area participation factors (apfs):

$apf_1 = 0.5$, $apf_2 = 0.25$, $apf_3 = 0.25$, $apf_4 = 0.5$, $apf_5 = 0.25$, $apf_6 = 0.25$. Coefficients that distribute ACE to several Gencos are termed as “ACE participation factors” (apfs). It should be noted that

$$\sum_{j=1}^m apf_j = 1, \text{ where } m \text{ is the number of Gencos. The}$$

Disco contract with the Gencos as per the following Disco participation matrix. The disco

participation matrix(DPM) in this work is taken as follows:

$$DPM = \begin{bmatrix} 0.1 & 0.0 & 0.3 & 0.4 \\ 0.0 & 0.1 & 0.0 & 0.2 \\ 0.3 & 0.4 & 0.1 & 0.0 \\ 0.2 & 0.0 & 0.2 & 0.1 \\ 0.2 & 0.3 & 0.0 & 0.1 \\ 0.2 & 0.2 & 0.4 & 0.2 \end{bmatrix}$$

A step load disturbance of 0.04 pu MW is considered in either of the areas (figure 7-9). In contract violation an additional load of 0.03 pu MW is considered in both the areas after the time span of 30 sec and 75 sec (figure 10-12). In this contract violation the Gencos which are present in that particular area where the violation has taken place indeed take up that extra load while the remaining generators of other area generate the power which they had been generating before.

Table1: Comparison of system performance with and without Fuzzy Logic controller

	Thermal area			Hydro area		
	Peak time (sec)	Overshoot	Settling Time (sec)	Peak time (sec)	Overshoot	Settling Time (sec)
Integral controller	0.375	0.0023392	0.615	0.31	0.0062918	1.815
Fuzzy Logic controller	0.335	0.0021661	0.485	0.3	0.0061976	1.665
% Improvement	10.66	7.3	21.13	3.2	1.49	8.2



Where % Improvement is given by

$$\left(\frac{(|\text{corresponding parameter in Integral controller}| - |\text{corresponding parameter in Fuzzy logic controller}|)}{|\text{Corresponding parameter in integral controller}|} \right) \times 100$$

Table 2: Comparison of Performance Index Values

	Performance Index Value
Integral controller	1.435×10^{-5}
Fuzzy Logic controller	8.068×10^{-6}

Figure 13 shows the comparison of performance index of the system during the base case and contract violation. Table 1 shows the comparison of dynamic performance of the system with and without fuzzy logic controller. Table 2 shows the performance index of the system during the base case. It can be observed from the tables that the system with fuzzy logic controller has better dynamic performance over the system with integral controller.

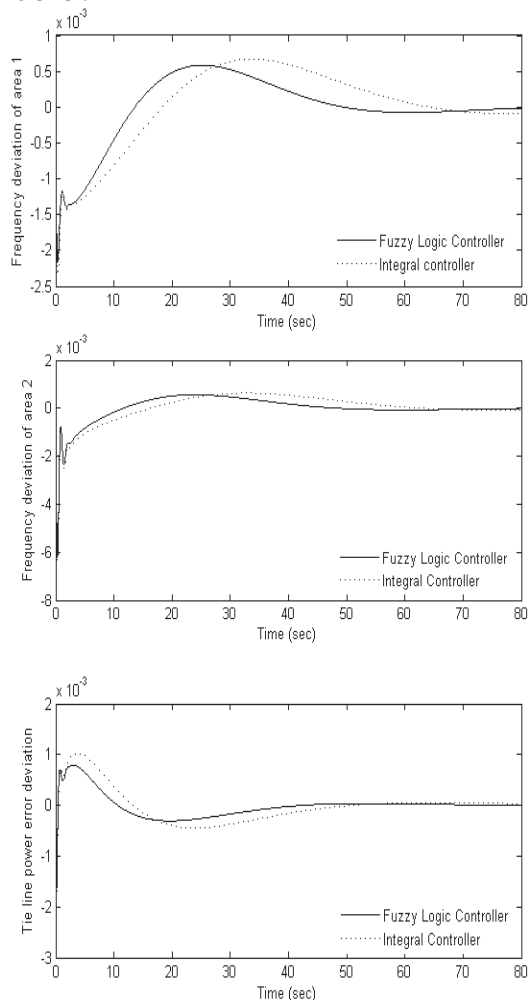


Figure 7: Frequency and tie line power deviations during normal case

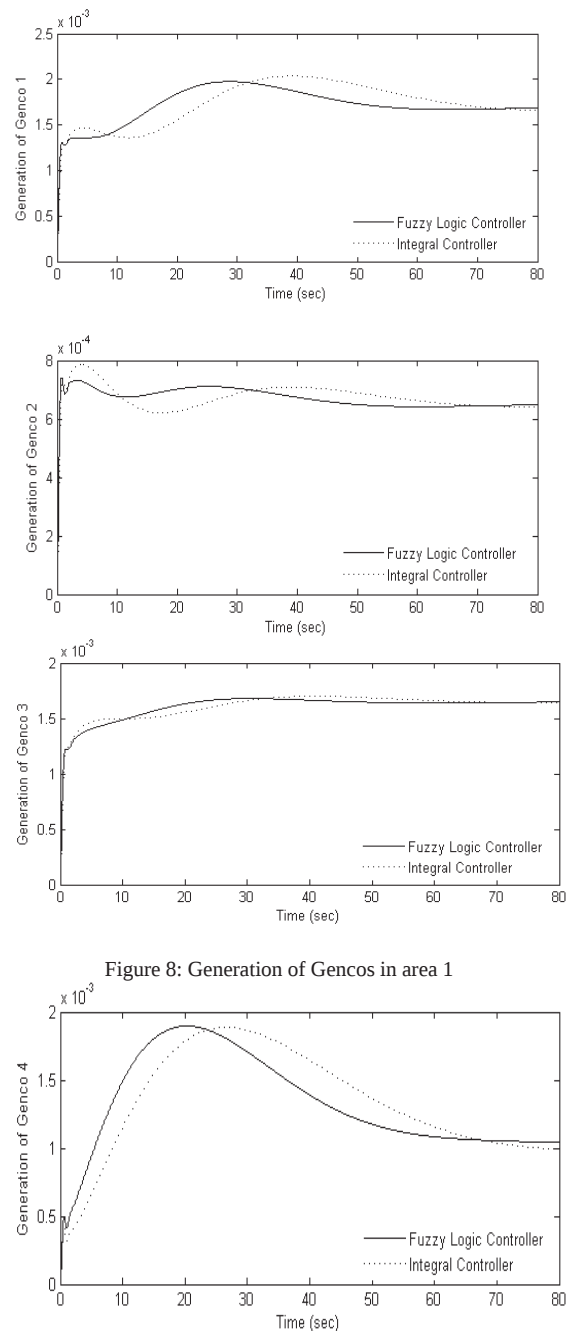
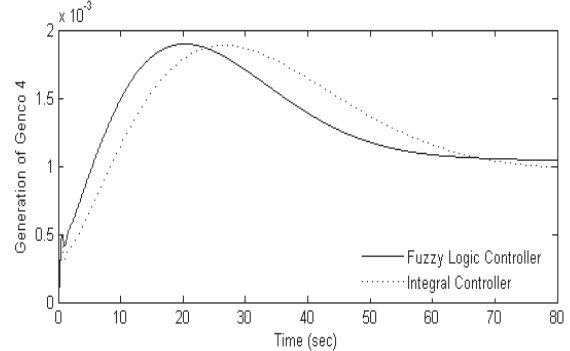


Figure 8: Generation of Gencos in area 1



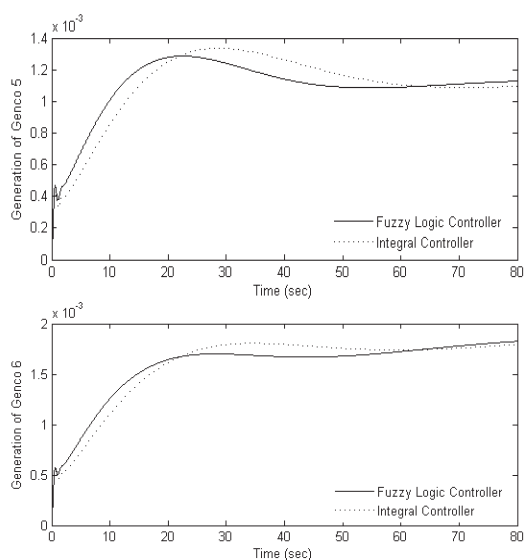


Figure 9: Generation of Gencos in area 2

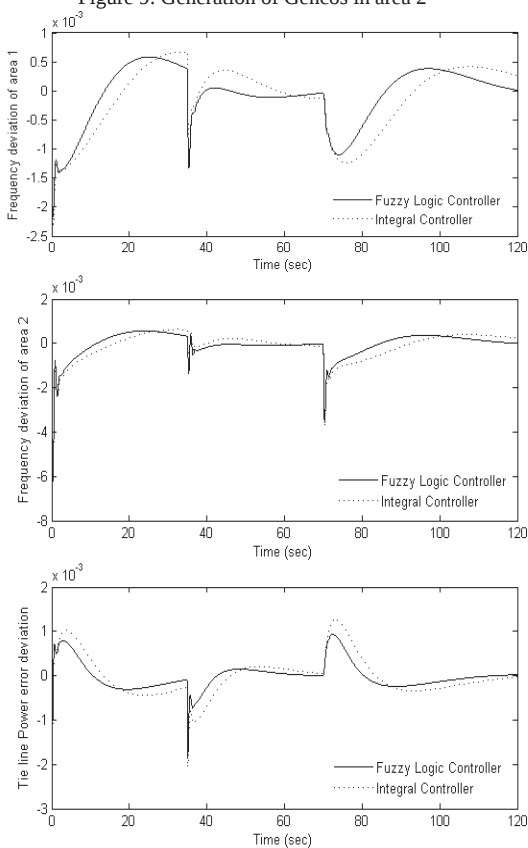


Figure 10: Frequency and tie line power deviations during contract violation case

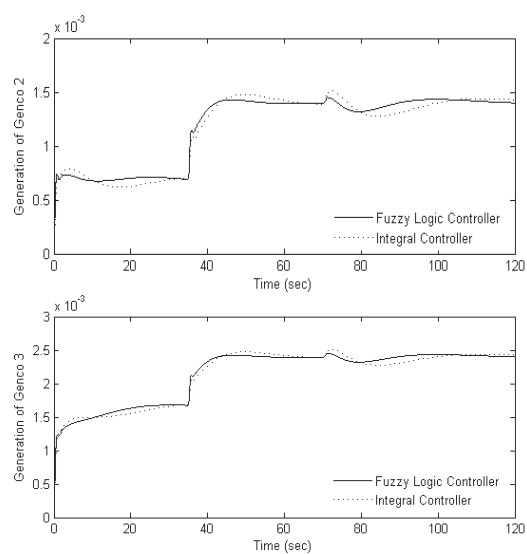
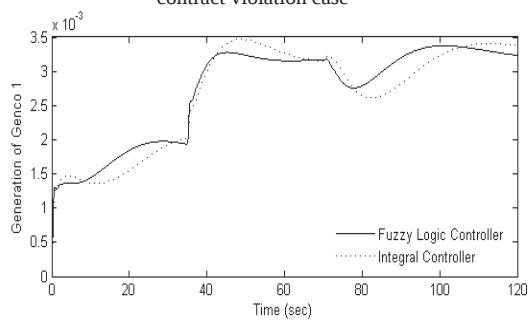


Figure 11: Generation of Gencos in area 1 during contract violation case

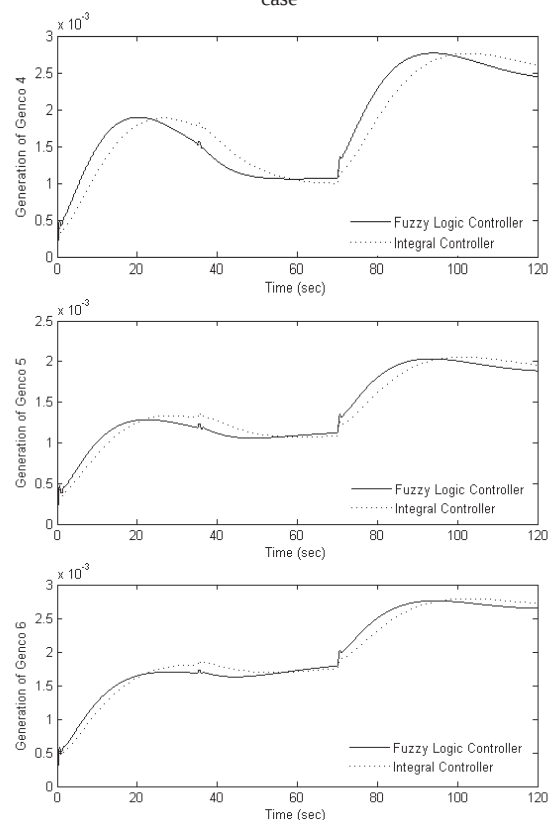


Fig 12: Generation of Gencos in area 2 during contract violation case

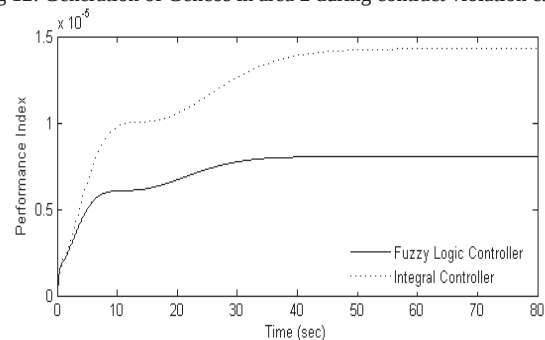


Fig 13: Comparison of performance Index



6. Conclusions

This paper has investigated the performance of conventional integral controller and Fuzzy Logic based controller on a two area hydrothermal system with SSSC and TCPS under deregulated scenario. The conventional design approach requires a deep understanding of the system, exact mathematical models, and precise numerical values whereas a basic feature of the fuzzy logic controller is that a process can be controlled without the knowledge of its underlying dynamics. This work builds a fuzzy rule base with the use of the area control error and rate of change of the error. Also a tie-line power flow control technique by SSSC and TCPS has been proposed for a two-area interconnected hydrothermal power system under open market scenario. The simulation results showed that the proposed fuzzy logic based controller yields more improved control performance than the conventional integral controller in the presence of TCPS and SSSC.

Appendix

All the notations carry the usual meanings

(a) System data

$T_{p1}, T_{p2} = 20s$; $K_{p1}, K_{p2} = 120 \text{ Hz/p.u.Mw}$;
 $P_{r1}, P_{r2} = 1200 \text{ Mw}$; $T_t = 0.3s$; $T_g = 0.08s$, $T_w = 1s$;
 $T_r = 5s$, $T_1 = 41.6s$, $T_2 = 0.513s$; $R_1, R_2 = 2.4\text{Hz/pu Mw}$;
 $T_{12} = 0.0866s$; $B_1, B_2 = 0.4249\text{p.u Mw/Hz}$;
 $T_1 = 0.279$; $T_2 = 0.026$; $T_3 = 0.411$; $T_4 = 0.1$;
 $K_{SSSC} = 0.1808$; $T_{SSSC} = 0.0386$

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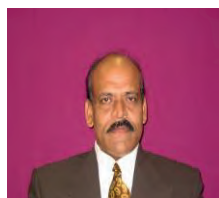
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Biographies



fuzzy logic, deregulated market

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An Efficient Variable Delay Random PWM Algorithm based Direct Torque Controlled Induction Motor Drive for Noise Reduction

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Abstract

Though the deterministic SVPWM algorithm gives good performance, it generates more acoustical noise. Hence, to mitigate the acoustical noise of the drive, this paper presents an efficient variable delay random pulse width modulation (VDRPWM) algorithm for direct torque controlled induction motor drive. The proposed VDRPWM technique is developed by using the notion of imaginary switching times, which are proportional to the instantaneous phase voltages. Hence, the proposed algorithm did not use angle and sector information and hence reduces the computational effort. Hence, the implementation of the proposed algorithm is simpler. Moreover, the proposed algorithm randomizes the switching periods by varying the delay of switching cycles with respect to corresponding sampling cycles. This randomization of switching periods has a mitigating effect of the acoustic and electromagnetic noise emitted by the supplied system. To validate the proposed PWM algorithm, simulation studies have been carried out and results are presented. The simulation results confirmed the features of proposed PWM algorithm.

Keywords: acoustic noise, direct torque control, imaginary switching times, space vector pulsewidth modulation (SVPWM), variable delay random pulsewidth modulation (VDRPWM).

1. Nomenclature

v_{ds}	d-axis stator voltage, V
v_{qs}	q-axis stator voltage, V
i_{ds}	d-axis stator current, A
i_{qs}	q-axis stator current, A
ψ_{ds}	d-axis stator flux linkage, wb-turn
ψ_{qs}	q-axis stator flux linkage, wb-turn
R_s	Stator resistance, ohm
R_r	Rotor resistance, ohm
L_s	Stator inductance, H

L_r	Rotor inductance, H
L_m	Mutual inductance, H
P	Number of poles
J	Moment of inertia, Kg-m ²
v_{ds}^*	d-axis reference stator voltage, V
v_{qs}^*	q-axis reference stator voltage, V
V_{dc}	DC link voltage, V
V_{an}, V_{bn}, V_{cn}	3-phase instantaneous phase voltages, V
V_{ref}	Magnitude of the reference voltage vector, V
α	Angle of the sampled reference voltage vector measured from the starting boundary of the given sector
θ_{sl}	slip angle
T_s	Duration of subcycle or sampling period, sec
T_1	Time duration of first voltage vector, s
T_2	Time duration of second voltage vector, s
T_z	Time duration of zero voltage vectors, s
M_i	Modulation index
T_{an}, T_{bn}, T_{cn}	Imaginary switching time periods proportional to the instantaneous values of the 3-phase reference voltages

2. Introduction

Research interest in high-performance control algorithms for induction motor received researcher's attention extensively over the last three decades. The invention of field oriented control (FOC) algorithm by F. Blaschke has brought renaissance in the high-performance applications. Though FOC gives good performance, the complexity involved is more due to the reference frame transformations. Moreover, the torque is also controlled in indirect manner in FOC technique. To overcome the drawbacks of FOC, a new control strategy renowned as direct torque control (DTC) was developed by Takahashi



[1]. Due to the simplicity, it has more applications in the industry [2]. A detailed comparison between FOC and DTC has given in [3]. Though conventional DTC (CDTC) is simple and easy to implement, it generates substantial steady state ripples in torque and current. Moreover, CDTC gives variable switching frequency operation of the inverter.

To reduce the steady state ripple and to get the constant switching frequency, several pulsewidth modulation (PWM) techniques were proposed to DTC algorithm. A detailed survey of the various PWM algorithms is given in [4]. Voltage control in PWM-inverter is performed by means of enforcing appropriate duty ratios of the inverter switches. The duty ratio, which is the ratio of the ON-time of the switch to the length of the switching (sampling) interval, does not depend on the location on the ON interval within the switching interval, i.e., the pulse position, or on the length of the switching interval, i.e., the switching frequency. The PWM techniques, which are given in [4] are also known as deterministic PWM algorithms. Among the various PWM algorithms, space vector PWM (SVPWM) algorithm is attracting many researchers due to its advantages. The SVPWM algorithm results in higher line side voltage and less line current harmonic distortion than sine PWM algorithm with constant switching frequency [5]. To reduce the steady state ripple and to get constant switching frequency operation of the inverter, SVPWM algorithm is applied to DTC in [6-7]. Though, the SVPWM algorithm gives good performance, it gives more acoustical noise and harmonic distortion due to the deterministic nature of pulse durations.

If either the pulse position or the switching frequency is varied in a random manner, the power spectrum of the output voltage of the converter acquires a continuous part, while the discrete (harmonic) part is significantly reduced. This is the basic principle of the random pulse width modulation (RPWM) which has in recent years attracted the increasing interest of researchers. The detailed review of the RPWM algorithms is given in [8]-[9]. Among, various RPWM algorithms, random pulse position PWM algorithms are easier for implementation [10]-[11]. However, a novel algorithm known as variable delay RPWM (VDRPWM) is reported recently [12]-[14]. The VDRPWM algorithm is characterized by a constant switching frequency and a varying switching period (T_s) realized by random changes of the delay of switching cycles with respect to the corresponding sampling cycles. However, the existing VDRPWM algorithm requires angle and sector information, which increases the complexity involved in the algorithm. To reduce the complexity involved in the conventional space vector approach, various PWM algorithms have been developed in [15]-[16] by using the concept of imaginary switching times.

This paper presents a simplified VDRPWM algorithm for direct torque controlled induction motor drive by using the concept of imaginary switching times for reduce acoustical noise and harmonic distortion.

The remainder of the paper is organized as follows: Section (2) focuses on classical SVPWM algorithm, Section (3) emphasizes on simplified SVPWM algorithm, Section (4) focuses on simplified VDRPWM algorithm,

Section (5) focuses on proposed VDRPWM based DTC-Im drives, Section (6) emphasizes on simulation results and discussion and the conclusions drawn are given in Section (7).

3. Classical SVPWM Algorithm

The three-phase, two-level voltage source inverter (VSI) has a simple structure and generates a low-frequency output voltage with controllable amplitude and frequency by programming high-frequency gating pulses. For a 3-phase, two-level VSI, there are eight possible voltage vectors, which can be represented in the space as shown in Figure 1.

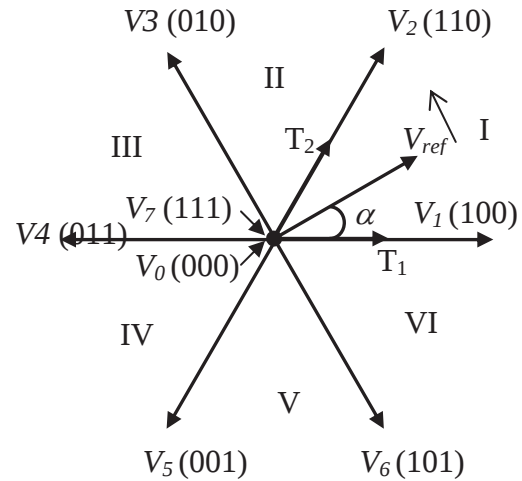


Figure 1 Possible voltage space vectors and sector definition

The V_1 to V_6 vectors are known as active voltage vectors and the remaining two vectors are known as zero voltage vectors. The reference voltage space vector (V_{ref}) represents the desired value of the fundamental components for the output phase voltages. In the space vector approach V_{ref} can be constructed in an average manner. The reference voltage vector (V_{ref}) is sampled at equal intervals of time, T_s referred to as sampling time period. The possible voltage vectors that can be produced by the inverter are applied for different time durations within a sampling time period such that the average vector produced over the T_s is equal to V_{ref} , both in magnitude and angle. It has been established that the vectors to be used to generate any sample are the zero voltage vectors and the two active voltage vectors forming the boundary of the sector in which the sample lies. As all six sectors are symmetrical, the discussion is limited to the first sector only. For the required reference voltage vector, the active and zero voltage vectors times can be calculated as in (1) - (3).

$$T_1 = \frac{2\sqrt{3}}{\pi} M_i \sin(60^\circ - \alpha) T_s \quad (1)$$

$$T_2 = \frac{2\sqrt{3}}{\pi} M_i \sin(\alpha) T_s \quad (2)$$

$$T_z = T_s - T_1 - T_2 \quad (3)$$



where M_i is the modulation index and defined as $M_i = \pi V_{\text{ref}} / 2V_{dc}$. In the SVPWM algorithm, the total zero voltage vector time is equally divided between V_0 and V_7 and distributed symmetrically at the start and end of the each sampling time period. As the classical SVPWM algorithm gives the switching times in a deterministic manner, it is also known as deterministic SVPWM algorithm.

4. Simplified SVPWM Algorithm

As the classical SVPWM algorithm uses angle and sector information for the calculation of switching times, the complexity involved in the algorithm is more. To reduce the complexity involved in the classical SVPWM algorithm, this paper presents a simplified SVPWM algorithm, which uses the concept of imaginary switching times. The imaginary switching time periods, which are proportional to the instantaneous values of the reference phase voltages, are defined as given in (4) [15]-[16].

$$T_{an} = \frac{T_s}{V_{dc}} V_{an}; T_{bn} = \frac{T_s}{V_{dc}} V_{bn}; T_{cn} = \frac{T_s}{V_{dc}} V_{cn} \quad (4)$$

Then, in each sampling time period, the maximum, minimum and middle values of imaginary switching times are evaluated by using (5)-(7).

$$T_{\max} = \text{Max}(T_{an}, T_{bn}, T_{cn}) \quad (5)$$

$$T_{\min} = \text{Min}(T_{an}, T_{bn}, T_{cn}) \quad (6)$$

$$T_{\text{mid}} = \text{Mid}(T_{an}, T_{bn}, T_{cn}) \quad (7)$$

Then the active voltage vector switching times T_1 and T_2 may be expressed as [16]

$$T_1 = T_{\max} - T_{\text{mid}}; T_2 = T_{\text{mid}} - T_{\min} \quad (8)$$

The zero voltage vectors switching time is calculated as

$$T_z = T_s - T_1 - T_2 \quad (9)$$

By using (8) and (9), the active vector times and zero vector times can be calculated. Thus, the active state times and zero states times can be calculate without determining the angle and sector information with the help of imaginary switching times.

5. Proposed VDRPWM Algorithm

The ideal RPWM algorithm should possess reduced acoustic noise over the full operating range, easier implementation, reduced complexity, acceptable switching losses and minimal impact on the system dynamics and basic motor control functionality. The RPWM algorithms can be classified into two categories as random switching frequency PWM (RSFPWM) and fixed switching frequency PWM algorithms. In the RSFPWM algorithms, both the sampling and PWM periods are synchronized. The major drawback of the RSFPWM algorithm is the limitation of the maximum code size by the minimum sampling time period. The fixed switching frequency RPWM algorithms can be classified as random zero state distribution PWM (RZSDPWM), random center distribution PWM (RCDPWM) and random lead-lag PWM (RLLPWM) algorithms. Though the fixed switching frequency RPWM algorithms allow optimal use of the processor computational capability due to fixed sample rate, these suffer from few limitations. RZSDPWM and RCDPWM algorithms lose effectiveness at higher modulation

indices. The RLLPWM algorithm does not offer a very good performance with respect to the reduction of acoustic/EMI emissions and suffers an increased current ripple as well. Additionally, both RLLPWM and RCDPWM introduce an error in the fundamental component of current due to a per-cycle average value of the switching ripple. Hence, VDRPWM algorithm is used in this paper for direct torque controlled induction motor drive. However, the existing VDRPWM algorithm uses sector and angle information for the calculation of gating times, which increase the complexity. Hence, to reduce the complexity involved in the algorithm, the concept of imaginary switching times is used and the gating times are calculated as explained in the previous section.

The proposed simplified VDRPWM algorithms uses fixed sample rate for optimal usage of processor computational power while providing quasi-random PWM output for good spectral spreading. This algorithm introduces a random delay into the trailing edge of the next PWM output cycle. Since two consecutive edges determine the PWM output period, a quasi-random PWM output is created as shown in Figure 2. The detailed flowchart of the proposed VDRPWM algorithm is shown in Figure 3, which illustrates the computation of the delay and switching period within the processor. In the proposed VDRPWM algorithm, a random number is generated between 0 and 1. Then, this is multiplied by sample time period to obtain the random delay time. Hence, the delay time is varied from zero to sampling time period. To avoid very short output PWM periods, a lower limit on the switching period is defined as minimum sampling time period (T_{swmin}). If the initial switching period calculation is less than the T_{swmin} , it will be clamped to T_{swmin} . Hence, the final delay must be recalculated after the limiting function. Thus, the resultant switching period may vary from T_{swmin} to 2 times the sampling time period. The average switching period will equal the sample period over time.

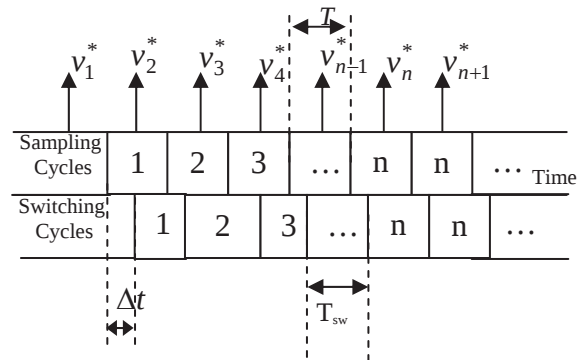


Figure 2 sampling and switching cycles in the proposed VDRPWM algorithm

The VDRPWM does not affect the basic motor control algorithm or type of space vector modulation employed. Thus, the VDRPWM algorithm consists of two steps. First, the duty cycles are computed just as SVPWM algorithm with fixed frequency. In second step, the delay is now computed as well. Then the duty cycles and delay are passed to the PWM modulator. Thus, the VDRPWM can be added to the inverter with minimal impact on the motor control algorithm. The switching loss of the



inverter is also same as SVPWM algorithm. Moreover, as it uses fixed sample rate, it avoids the updating of filter and regulator gains, which is necessary when using variable sample rate techniques.

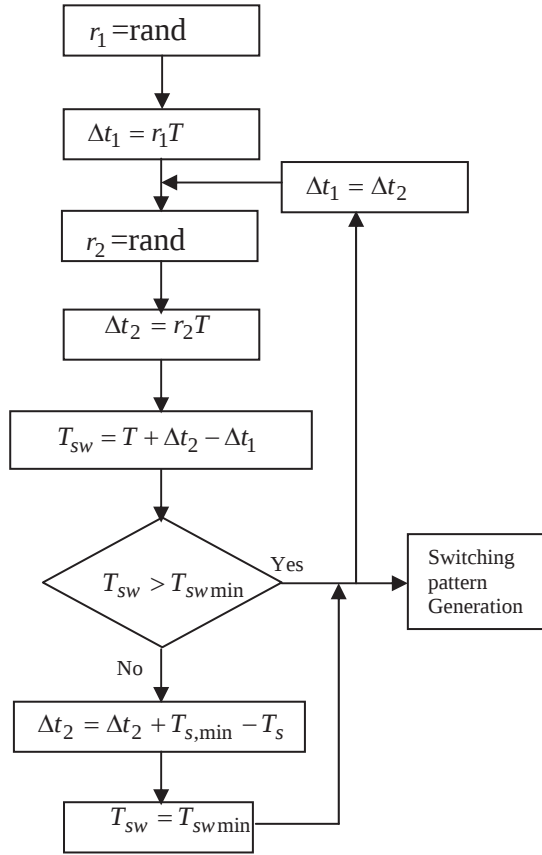


Figure 3 flowchart of the proposed VDRPWM algorithm

6. Proposed VDRPWM Algorithm Based DTC-IM Drive

The reference phase voltages can be obtained in many ways. But, to reduce the complexity of the algorithm, in this paper, the required reference voltages are constructed by using the errors between the reference d-axis and q-axis stator fluxes and d-axis and q-axis estimated stator fluxes sampled from the previous cycle. The block diagram of the proposed VDRPWM algorithm based DTC is as shown in Fig. 4. From Fig. 4, it is seen that the proposed VDRPWM based DTC scheme retains all the advantages of the CDTC, such as no co-ordinate transformation and robust to motor parameters. However a PWM modulator is used to generate the pulses for the inverter, therefore the complexity is increased in comparison with the CDTC method. In the proposed method, the position of the reference stator flux vector $\bar{\psi}_s^*$ is derived by the addition of slip speed and actual rotor speed. The actual synchronous speed of the stator flux vector $\bar{\psi}_s$ is calculated from the adaptive motor model. After each sampling interval, actual stator flux vector $\bar{\psi}_s$ is corrected by the error and it tries to attain the reference flux space vector $\bar{\psi}_s^*$. Thus the flux error is minimized in each sampling interval. The d-axis and q-

axis components of the reference voltage vector can be obtained as follows:

Reference values of the d-axis and q- axis stator fluxes and actual values of the d-axis and q-axis stator fluxes are compared in the reference voltage vector calculator block and hence the errors in the d-axis and q-axis stator flux vectors are obtained as in (10)-(11).

$$\Delta\psi_{ds} = \psi_{ds}^* - \psi_{ds} \quad (10)$$

$$\Delta\psi_{qs} = \psi_{qs}^* - \psi_{qs} \quad (11)$$

The knowledge of flux error and stator ohmic drop allows the determination of appropriate reference voltage space vectors as given in (12)-(13).

$$V_{ds}^* = R_s i_{ds} + \frac{\Delta\psi_{ds}}{T_{sw}} \quad (12)$$

$$V_{qs}^* = R_s i_{qs} + \frac{\Delta\psi_{qs}}{T_{sw}} \quad (13)$$

where, T_{sw} is the duration of subcycle or sampling period and it is a half of period of the switching frequency. This implies that the torque and flux are controlled twice per switching cycle. Further, these d-q components of the reference voltage vector are fed to the PWM block. In PWM block, these two-phase voltages then converter into three-phase voltages. Then, the switching times are calculated as explained in previous sections.

7. Simulation Results and Discussion

To verify the proposed scheme, numerical simulation studies have been carried out by using Matlab/Simulink. For the simulation, the reference flux is taken as 1wb and starting torque is limited to 15 N-m. The induction motor used in this case study is a 1.5 kW, 1440 rpm, 4-pole, 3-phase induction motor having the following parameters: $R_s = 7.83\Omega$, $R_r = 7.55\Omega$, $L_s = 0.4751H$, $L_r = 0.4751H$, $L_m = 0.4535H$ and $J = 0.06 \text{ Kg.m}^2$

The steady state simulation results of conventional DTC and SVPWM algorithm based DTC are shown in Figure 5 – Figure 8. From the simulation results, it can be observed that the total harmonic distortion (THD) of conventional DTC algorithm is very high. To reduce the THD of conventional DTC algorithm, SVPWM algorithm is used for DTC. From Figure 6 and Figure 8, it can be observed that the THD is less when compared with the conventional DTC algorithm. As the amplitudes of dominating harmonics around switching frequency are high in classical SVPWM algorithm, it gives more acoustical noise and harmonic distortion. To reduce acoustical noise, the simplified VDRPWM algorithm is proposed in this paper. The simulation results of proposed VDRPWM algorithm based DTC drive are shown from Figure 9 to Figure 14. From the simulation results, it can be observed that the proposed VDRPWM algorithm gives reduced THD when compared with the SVPWM algorithm. Moreover, as the amplitude of dominating harmonics is less when compared with the classical SVPWM algorithm, the proposed VDRPWM algorithm gives less acoustical noise. Moreover, the proposed VDRPWM algorithm gives spread spectra when compared with the SVPWM algorithm.



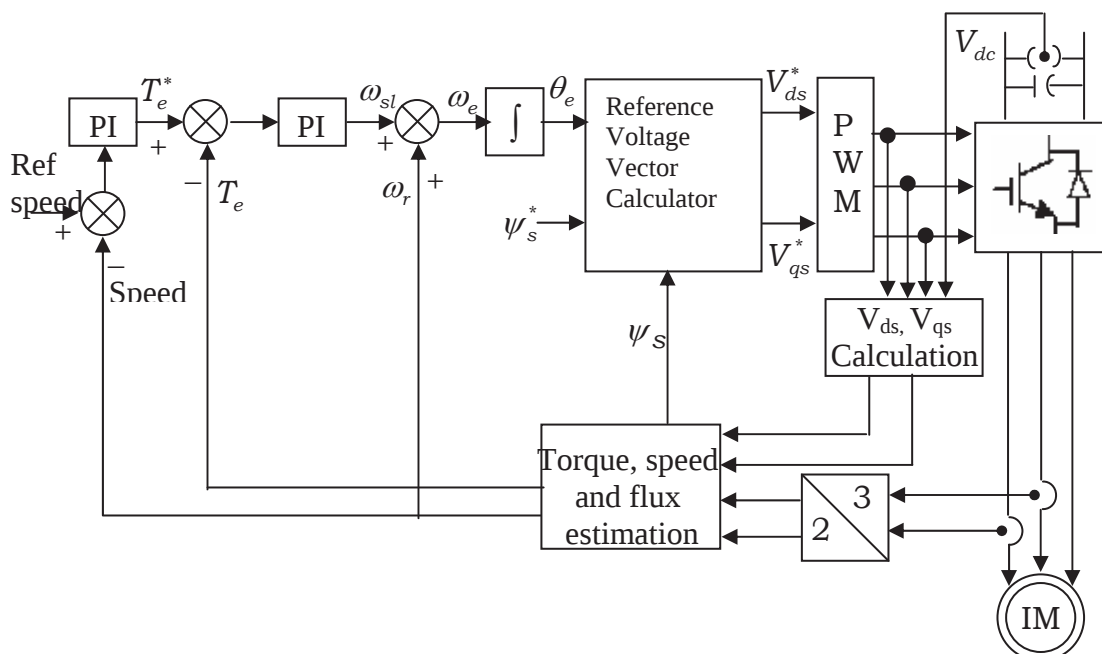


Figure 4. Block diagram of proposed VDZPWM based DTC

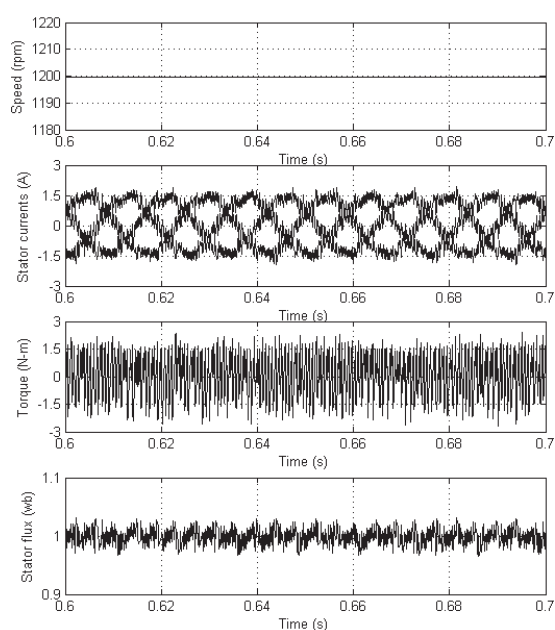


Figure 5 steady state plots of conventional DTC

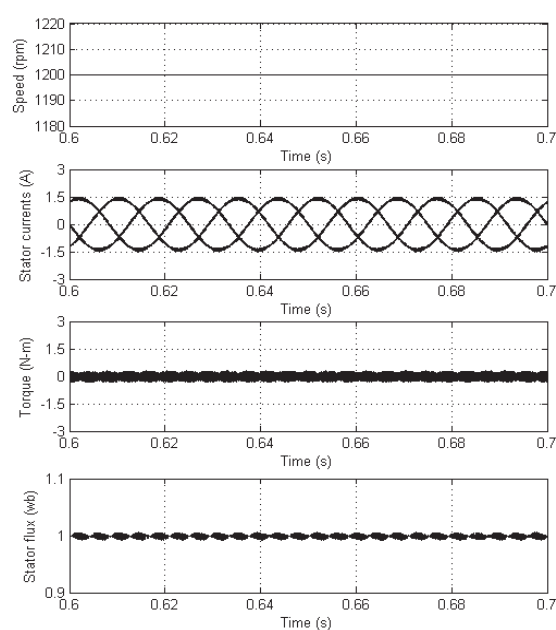


Figure 7 steady state plots of SVPWM algorithm based DTC

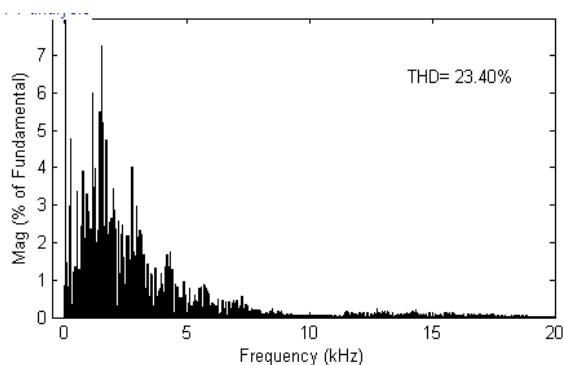


Figure 6 Harmonic spectra of line current in conventional DTC

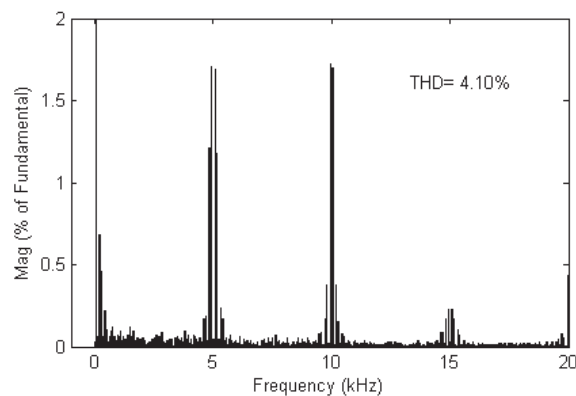


Figure 8 Harmonic spectra of line current in SVPWM based DTC



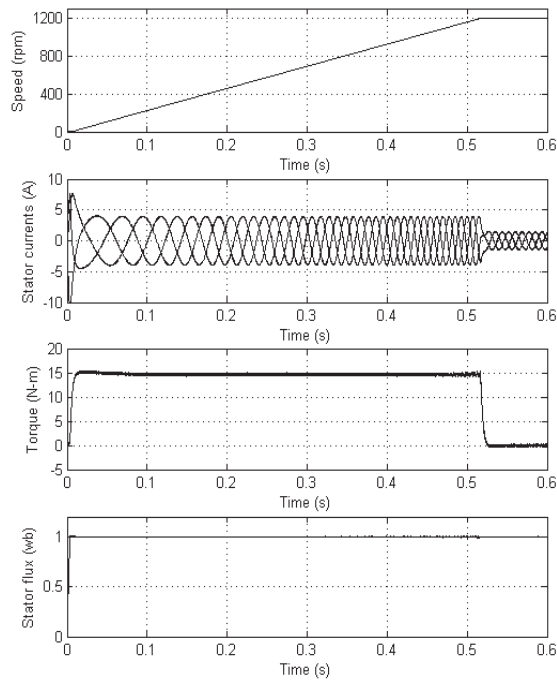


Figure 9 starting transients in proposed VDRPWM based DTC

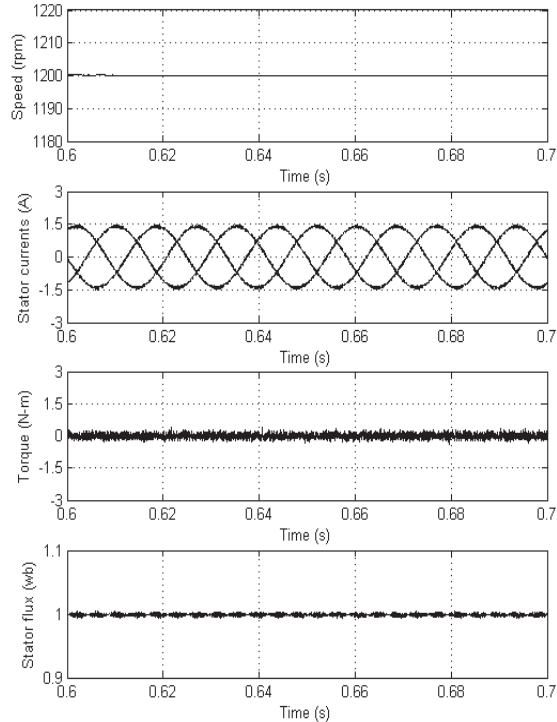


Figure 10 steady state plots of proposed VDRPWM based DTC

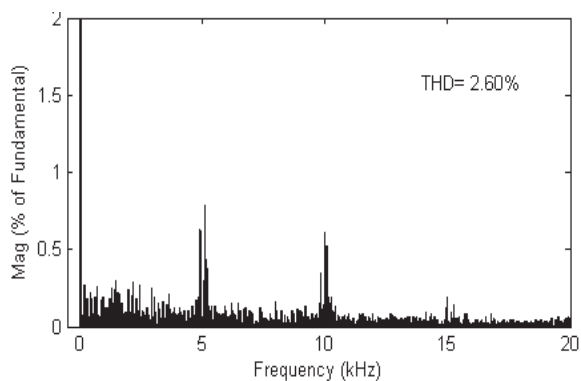


Figure 11 Harmonic spectra of line current in proposed VDRPWM based DTC

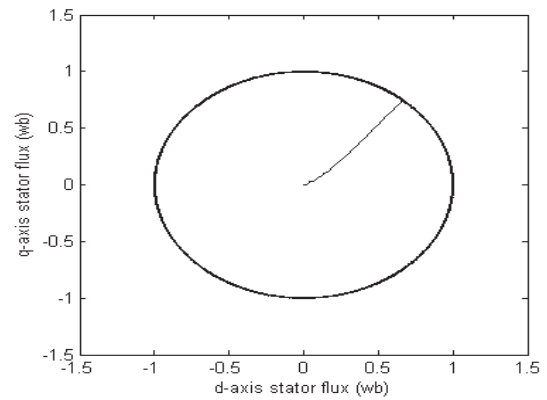


Figure 12 Locus of stator flux in proposed VDRPWM based DTC

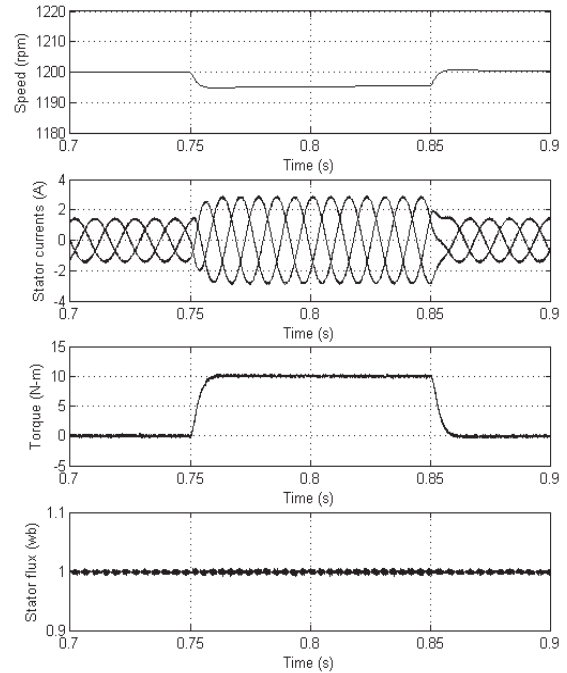


Figure 13 Transients during step change in load for proposed VDRPWM algorithm based DTC drive (a load torque of 10 N-m is applied at 0.75 s and removed at 0.85s)

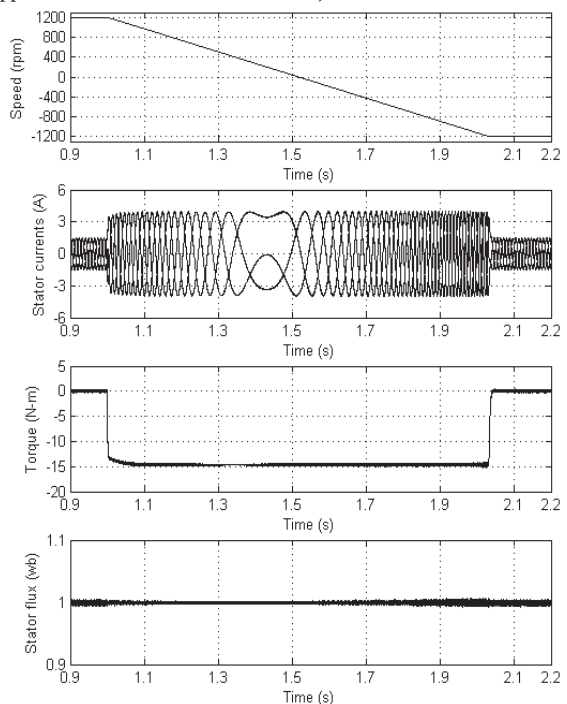


Figure 14 Transients during speed reversal condition for proposed VDRPWM algorithm based DTC drive



8. Conclusion

Though the SVPWM based direct torque controlled induction motor drive gives good performance, it generates more acoustical noise and harmonic distortion due to the dominating harmonics. Hence, to reduce the harmonic distortion and acoustical noise of the drive, simplified VDRPWM algorithm is proposed for direct torque controlled induction motor. As the proposed algorithm uses the concept of imaginary switching times, it reduces the complexity involved in the algorithm. From the simulation results, it can be observed that the proposed VDRPWM algorithm gives less harmonic distortion. As the magnitude of dominant harmonics around the switching frequency (5 kHz) is less in the proposed VDRPWM algorithm, it gives less acoustical noise when compared with the SVPWM algorithm.

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